APPENDIX 11

Analysis of USADR IBOC DAR System Modifications and Their Impact on Performance

Prepared for

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EXECUTIVE SUMMARY

USADR has proposed changes in their direct sequence spread spectrum IBOC system (FM-1) that claim to have solved problems in their original system. For this study, MDS developed a detailed computer model of the original FM-1 system and the modified system, using the best available information. From these computer simulations, we have determined that by decreasing "processing gain" (the relative amount of spectrum spreading of the digital signal), lowering injection level and increasing complexity, the bit error rate of the modified system went up and bandwidth went down. These changes have seriously impaired performance, however, and make the modified system much more susceptible to interference and multipath. Thus, in the modified system, there is much less chance the digital signal can be recovered successfully.

I. IBOC Report

A. Background

The USADR FM-1 DAB system uses Direct Sequence Spread Spectrum (DSSS) to transmit digital audio on specially shaped sidebands centered at approximately 180kHz on either side of the main channel analog modulation. There are fundamental concerns in operating a direct sequence system in so closely spaced in frequency to an analog FM system. At issue is interference - the IBOC signal can interfere with the analog FM signal or with other IBOC signals; likewise, the analog FM signal can interfere with the IBOC signal. In addition, the IBOC signal must contend with the issues that are systemic to operation in the FM band, such as multipath, co- and adjacent channel interference.

By their own admission, the original USADR FM-1 proposal exhibited some significant problems in these areas; these problems were listed as: [13]

1. DAB interference to host FM signal.
2. Interference to DAB from the first adjacent FM signal.
3. Interference to the FM signal from the first adjacent DAB.
4. Interference between DAB second adjacents.
5. Robustness of DAB in multipath fading environment."

In reviewing the available literature for these IBOC DAB systems, it became clear that, while extensive testing has been done of proposed IBOC systems, no detailed modeling effort has been performed independently in an attempt to quantify the problem areas above. In this report, Mobile Data Systems has developed a Matlab simulation of the FM-1 system, both original and proposed, in an attempt to evaluate the tradeoffs in system design. Modeling complex systems instead of designing and building hardware is an effective and important technique for evaluating high risk technology without the enormous expense of creating and testing hardware. It is somewhat surprising that a detailed simulation was not performed on the USADR system; even this modest effort does not include a complete simulation of the analog FM signal.

B. Modeling Issues

In order to model the FM-1 system, the design parameters must be known or inferred; while details of the original FM-1 proposal are well known, details on the proposed modifications are not as well documented in [13]. In fact, [13] also includes a brief description of an orthogonal frequency division multiplexing (OFDM) system for DAB that appears to be completely unrelated to the DSSS system; in this report, we will only model the modified DSSS system. The modified system is, at this time, only a proposal.

The system parameters, whether gathered from documentation, derived, or inferred, are listed in Appendix C. From these parameters, both versions of the FM-1 system were simulated, to include bit error rate (BER) as a function of the energy per bit (E_b) divided by the noise power spectral density (N_0). This ratio, E_b / N_0, is similar to signal to noise ratio commonly used in analog systems. In this case, N_0 can include effects such as adjacent DAB channel interference and other sources of "noise like" interference. The effect of BER should be obvious; as the number of bit errors increases, the performance of the system completely breaks down. It is characteristic of digital systems (such as digital cellular) to have a "wall" or critical BER at which the system becomes unusable. Thus, unlike an analog system where the user will hear a gradual degradation of the signal, a digital audio system will degrade rapidly and fail to operate.

It should be noted that models such as this can be most effectively used to compare the performance of the system to the theoretical limits; the simulation output will establish whether the system can work in the ideal case, but the results of testing with prototype systems will not be explicitly compared here; the reader is referred to [14], which is a comprehensive report of laboratory testing. The models can effectively establish whether a system could work under ideal conditions; since real world systems do not consist of ideal components, these simulations establish upper bounds on performance.

C. Comments and Conclusion

The issues addressed in the modeling efforts allow us to draw conclusions about the performance of the USADR systems, both the original and modified versions. In general, the conclusions are that the modified USADR system corrects some deficiencies in the original design, but worsens its performance in several critical areas, leading to the conclusion that the adjustment of system parameters is a technical "shell game"; the USADR system operates at the limits of direct sequence spread spectrum performance, and attempts to optimize one parameter will likely cause degradation in one or more other parameters. Only significant reductions in source coding rate can lead to any real improvement in system performance; apparently, USADR maintains the same program data rate in both system proposals.

One of the most significant findings in this study is that the modified USADR system has considerably less processing gain than the original system (23dB in original vs 11dB in modified), which makes it much more susceptible to multipath, co-channel interference from the analog signal, from itself, and from adjacent channel interference. Processing gain is a measure of the ability of a DSSS system to operate in the presence of interference, and higher is better. In the case of the modified system, since the processing gain is only 11 dB, the simulation shows a rather alarming bit error rate of 0.04 for the modified system, with no analog interference present. Thus, while the modified system fits within a narrower spectral mask, USADR had to lower the processing gain to do so.

We shall address several specific areas in light of the simulation results:

1. Host compatibility

In the modified system, the spreading bandwidth and processing gain was reduced, and the injection level was lowered 6-10dB in order to reduce the likelihood of interference to the analog program. Indeed, the modified system will certainly be less likely to cause interference, but the IBOC signal is much more fragile, and requires a far more complicated receiver.[1] Thus, the modified system is better in terms of digital interference on the analog host, but worse in interference to the DAB system.

2. Multipath

Of particular interest is the multipath performance of the modified USADR system, particularly in light of the reduced processing gain. As outlined in Appendix D, the USADR modified operates at a marginal 0.04 BER in the <u>absence</u> of multipath due to self-interference. If multipath causes an additional degradation of 3 dB in E_b / N_0, then the

[1] See Appendix C; the modified system will require 512 correlators running in parallel, which is over 5 times more complex than the original FM-1 system.

BER will further decrease to approximately 0.06, which will be extremely difficult to overcome using error correction techniques. One possible technique to improve this is use of a RAKE receiver, as described in Appendix C; the coherence bandwidth[2] of fades in the FM band is within the range of the FM-1 bandwidth, which means that a RAKE receiver may not be useful for improving the performance of FM-1 in multipath; RAKE receivers cannot improve the signal if is completely wiped out by a fade, and the modified system will gain less from a RAKE receiver than the original system. The only techniques that will be useful for a relatively narrow signal such as this are frequency or spatial diversity. USADR employs frequency diversity by selecting either upper or lower sideband, which gives an overall E_b / N_0 improvement related to the statistical independence of the sidebands.

II. Appendices

Mobile Data Systems has designed a mathematical/functional system to match the original USADR FM-1 Noise Modulation and Coding (NOMAC) system, and FM-1 enhanced system based upon Gold codes. The assumptions we have used to develop computer simulation results for both systems will be clearly stated. We designed a canonical functional description in order that theoretical performance bounds are readily available; we compare simulated results to theoretical bounds to ensure the simulation is accurate, thus achieving independence from details of specific hardware implementation unavailable to MDS.

Computer simulations predict a degraded performance of the 48 channel NOMAC FM-1 system due to loss of signalling orthogonality over finite length waveforms, and a CDMA-like self interference due to the overlay of 48 channels to increase the channel bit rate. Details of signal filtering to form the baseband spectrum are unknown.

A 32 channel, 32-ary, 64 bit Gold code based enhanced FM-1 system was simulated in order to estimate the baseband spectral content of the waveform suite and predict interference effects; raised cosine baseband filtering, and 32-ary biorthogonal signalling are assumed. The extent of RF filtering is unknown. Additionally, computer simulations were run to predict performance degradation from the self interference of a 32 signal overlay that agree with large sample statistics CDMA BER estimates. The simulations indicate that baseband pulse filtering is critical to creating a digital sideband IBOC signal that meets FCC spectral mask requirements. The receiver complexity for the proposed enhanced system is high, requiring 512 correlators, each of which is a length matched to the chosen Gold code length.

[2] The coherence bandwidth of a radio channel is the bandwidth over which the statistical properties are highly correlated. It is related to the "delay spread" in the channel, which is the amount of time that reflected signals can be observed.

A. Brief Outline of USADR system.

FM-1 NOMAC System Assumptions

The digital communications signaling system for the basic FM-1 system presented in [13] is assumed to be a 48 channel noise modulation (NOMAC) signalling scheme, wherein each channel is binary with biorthogonal coding. Each signalling waveform is assumed to be 192 samples from a random number generator, subsequently band pass filtered to control the baseband spectrum. The details of the filter are in a patent application, as well as how the filtered waveforms are then orthogonalized to remove filter induced correlation.

In order to avoid these unknowns, our baseband computer simulation operates without the bandpass filtering so as to maintain maximum signal orthogonality. We could of course design a BPF to match the plots supplied in [14], however we would have no way to the orthogonalize the resultant 48 band pass filtered signalling waveform suite. In any event this conservative approach bounds the performance, which is already degraded from the optimum for reasons later discussed in this report.

If the bit rate per channel is 8 kbps the aggregate channel bit rate is 384 kbps. At these rates the random number sequence would require a sampling rate of 1.536 MHz. If a rate 1/2 Forward Error Correction (FEC) algorithm is utilized to guard against channel impairments, then a 192 kbps information data rate is achieved.

The simulated results are compared to mathematical theoretical results available for M-ary biorthogonal signalling to quantify the performance loss. We use theoretical canonical performance bounds so that we are insulated from any specific hardware implementation details, which are not provided. Additionally, use of theoretical m-ary bounds provides an absolute limit on the best performance that can be achieved; any real world system or simulated system can not exceed the theoretical bound, although systems without cost constraint may well be designed to operate near or at theoretical limits. The signal spectrum is not estimated as we can not accurately model the baseband filtering applied to the pseudo random number sequence used as the signalling suite. It may be noted, however, that clearly the noise waveforms can be filtered to meet a desired spectral mask, but the performance will degrade measurably unless the filtered signal suite is perfectly orthogonalized. Additionally, a 48 signal overlay suffers from the impairments common to any CDMA system in that a performance floor is set by the processing gain and the number of users, even if infinite signal to noise ratio (SNR) is assumed.

The receiver for NOMAC digital systems is straightforward, requiring 48 correlators, each operating on 192 waveform samples, an operation that can be achieved in real time DSP using FFT based high speed correlation algorithms, programmed into commercially available DSP chips.

FM-1 Enhanced System Using Gold Codes

The FM-1 enhanced system, based on a suite of 32 channel, 64 bit (assumed) Gold codes, and 32-ary signally, was simulated to estimate the spectral content of the transmitted signal. The processing gain of this spread spectrum system is directly proportional to the length of the Gold codes used; clearly the higher the processing gain the better multipath tolerance and narrow band interference rejection, but also the high chip rate that must be supported by the channel. The processing gain of any spread spectrum system, including this spread spectrum system, has absolutely no effect on performance in the presence of additive white Gaussian noise; this means that any interference from adjacent channel IBOC systems that appears as white Gaussian noise or statistically similar is not mitigated by increasing or decreasing processing gain. Processing gain for direct sequence spread spectrum (DSSS) systems is beneficial only to combating multipath and narrow band interference. The simulation is used to estimate performance degradation to the self interference effects of a 32 signal overlay without benefit of significant spectrum spreading and compared to theoretical bounds as a cross check. Specific hardware implementations will cause additional unavoidable losses in processing gain of as much as 3 dB.

As determined from [13], we assume availability of a 96 kbps stereo audio codec algorithm, a rate 1/2 FEC, and a bandwidth efficiency of at least 2 bps/Hz, so that the signal will fit the 100 kHz spectral mask set by the FCC. Therefore, each of the 32 channels must operate at a bit rate of 6 kbps to produce an aggregate channel bit rate of 192 kbps. We can constrain the baseband spectrum to 100 kHz by use of raised cosine pulse shape filtering (or any similar method such as Gaussian filtering in GMSK), which in turn constrains the per channel chip rate. For narrow band interference, self interference protection, and multipath robustness we desire the Gold code length to be as long as possible to increase the spread spectrum processing gain. The self interference protection is important because 32 signalling suites are transmitted together, thus the receiver sees the mathematical equivalent of a 32 user CDMA system with perfect power control. The length of the spreading code does not alter BER performance against Additive White Gaussian Noise (AWGN). If we set the Gold code length to 64 bits, then use of 32-ary biorthogonal signalling induces a chip rate of 76.8 kcps, a rate that with proper pulse filtering can operate within a 100 kHz bandwidth (baseband). The processing gain is $10\text{Log}(64/5) = 10\text{Log}(76.8 \text{ kcps}/6 \text{ kbps}) = 10\text{Log}(12.8) = 11$ dB.

Compared to the FM-1 NOMAC system the complexity is significantly increased. The receiver requires $(32)(16) = 512$ (assuming biorthogonal signalling) 64 bit correlators in order to decode the channel bits, then a rate 1/2 decoder to decode data bits. By contrast, FM-1 NOMAC required a set of 48 correlators, each 192 samples in length.

Biorthogonal signalling requires coherent demodulation as the correlators much detect the sign of the correlation pulse; if orthogonal signalling is used a noncoherent demodulator may be used, but the number of correlators doubles to 1024, assuming the system we have constructed.

B. Input Data

1. a) FM-1 NOMAC System

A pseudo random sequence length of 192 is assumed (not stated in USADR report) as the example autocorrelation plot was 384 samples in length since a discrete time autocorrelation estimate typically doubles the length of the discrete random data. The pseudo random sequence sample rate was stated in the USADR report as 1.536 MHz; assuming a 192 pseudo random number sequence produces a per channel rate of 8 kbps. If the aggregate channel rate is 384 kbps as stated in USADR report, then if each channel operates at 8 kbps it is required to overlay 48 channels. A pseudo random number sequence enjoys infinite bandwidth, thus baseband filtering is required to meet FCC spectral mask requirements; for obvious reasons the filtered pseudo random sequences must be re-orthogonalized since filtering a random sequence induces correlation between signal samples. The filtering is mentioned, but not specified, in the USADR report. The M-ary communication system theoretical performance assumes perfectly orthogonal signals, thus serves as a bound on achievable performance. Any realistic system can approach, but never exceed these bounds; such an analytical approach is independent of specific system implementation and bounds performance.

1. b) FM-1 Enhanced Gold Code System

The example autocorrelation and cross correlation plots in the USADR report indicated use of 127 bit Gold codes. The chip rate (ie, the rate the Gold code bits are. transmitted) was stated in the USADR report to be 75.6 kcps. It was further stated that 32-ary system could be constructed with a 32 channel overlay to meet the channel rate requirements stated to be 192 kbps.

If we assume a 32 channel system, with an aggregate channel rate of 192 kbps, then each channel must operate at 6 kbps. If we set the Gold code length to 63 bits, vice 127 bits, and use the stated 32-ary signalling, then the chip rate is (63/5)(6 kbps)=75.6 kcps, which matches the chip rate stated in the USADR report. A system constructed in this manner would exhibit a progressing gain of approximately 11 dB.

2. Parameters that are needed but not supplied by USADR.

a) Parameters Required To Analyze Digital Communication Systems

Since modern day digital communications systems usually require power efficiency and bandwidth efficiency as a primary design constraint, coherent digital modulation techniques are used extensively. Common coherent digital modulation signaling methods are binary phase shift keying (BPSK), quadrature phase shift keying (QPSK), minimum shift keying (MSK), Gaussian MSK (GMSK), M-ary PSK, M-ary Quadrature Amplitude Modulation (QAM), and orthogonal/biorthogonal signally (may be binary or M-ary). The modulation methods (BPSK, QPSK, MSK, and orthogonal/biorthogonal) provide the nearly the same BER performance, optimal against an additive white Gaussian noise (AWGN) channel without multipath, but differ in complexity and power spectra characteristics.

The bandwidth of a digitally modulated signal and the shape of the signal spectrum are determined by the type of digital modulation, type of baseband data encoding utilized, and the data rate. The data itself is commonly encoded as non-return to zero (NRZ) or Manchester (biphase) format, although other data encodings are possible. Additionally, it is a near certainty that the digital data will be subjected to source coding in order to reduce the baseband data rate. For example, digital broadcast audio data will generally be coded to a lower bit rate via routine parametric coding algorithms such as Code excited Linear Prediction (CELP), prior to modulation. It is common to combine source coding with some form of channel coding for error protection (the source coded bits with less redundancy are more sensitive to bit errors than digitized raw audio). A multipath communications channel is considered very harsh; thus, in order to ensure reliable communications, channel coding may be invoked to protect some or all of the source encoded bits. Although source coding decreases the data rate, channel coding increases the data rate, so the two coding strategies are at odds with each other. Both are usually necessary, however, so a digital system receiver designer must deal with the added complexity involved from a system design viewpoint. As might be expected, the higher complexity modulation types such as MSK offer greater bandwidth efficiency, thus allowing higher data rates to be accommodated for a given RF bandwidth. The trade-offs between data rate (a function of baseband data rate, source coding, and channel coding), modulation type, and processing complexity are virtually endless.

b) Parameter Assumptions - Line Coding

A line code maps the logical binary data to analog voltage levels for subsequent modulation onto a carrier and transmission over the ether. Common lines codes are Manchester, NRZ, and RZ, each possessing advantages and disadvantages. NRZ is the most bandwidth efficient and enjoys the lowest BER, but requires DC coupling, and performs the poorest within the bit synchronization algorithm in the digital receiver. Manchester coding possesses the same BER as NRZ, is easy to bit synchronize, and admits AC coupling, but requires double the bandwidth of NRZ. RZ is a compromise between NRZ and Manchester, trading off BER performance for AC coupling and bandwidth requirement somewhere between NRZ and Manchester.

The original NOMAC system does not use a line code as its a binary valued noise modulation system. The enhanced system was assumed to use NRZ line coding in order to minimize bandwidth.

c) Baseband Pulse Shaping

The originally proposed NOMAC system does not utilize binary pulses, rather uses pseudo random noise sequences as a signal suite. These sequences (infinite bandwidth) must be filtered to control the spectral content. The baseband filter parameters are unknown; in any event, if we constructed a filter we then could not re-orthogonalize the filtered sequences. As a consequence the filtering was not used so that our simulation operates in a more ideal manner than a real system and serves as a performance bound. Additionally, we can theoretically bound performance from M-ary detection theory, and compare our simulation to a mathematically perfect system.

The spectrum of rectangular pulses is infinite in extent with a strong sin(x)/x sidelobe structure, not suitable for any digital system that is required to operate in a limited spectral mask. Common pulse shaping filters to limit spectral occupancy and/or to mitigate intersymbol interference are Nyquist filtering, Gaussian filtering, and raised cosine filtering, although many other filters may be used. We chose arbitrarily a square root raised cosine baseband pulse filter as it is specified in the IS-54 North American Digital Cellular Standard. Such a filter may be designed to constrain the spectrum as tightly as desired, however, the receiver performance degrades as the bandwidth is decreased, as might be expected. We used a compromise value of beta-1.0 filter, which was the widest bandwidth that could just meet the spectral mask easily.

d) Data Rates

The digital data rate simply refers to the number of bits per second transmitted across the channel. The higher the data rate the wider the bandwidth all other factors being equal. We chose data rates to match data rates stated in the USADR report as previously discussed in this report.

e) Digital Modulator

Our simulation operates at baseband (view the bandpass spectrum as frequency shifted to baseband) as do all digital communication system simulations as it is not feasible to sample RF waveforms accurately. Imagine sampling an 88 MHz FM carrier at the Nyquist rate of 176 megasamples per second and trying to run the computer program. Baseband analysis is equivalent to bandpass RF analysis. We note, however, that the USADR system, assuming biorthogonal signalling, must implement a coherent modulator/demodulator. Furthermore, the baseband spectrum must translate (modulate) to the FM radio band without spectrum doubling so that DSB can not be used (unless chip rate is cut in half).

f) IF/RF Filtering

A digital communications transmitter requires IF and RF filtering to control spectral content. We possess no knowledge of the filtering applied. The simulations operate at baseband and do not include additional filtering, thus we provide a performance bound.

APPENDIX C. The model

The model simulation was coded in Matlab, a commonly available software package based on linear algebraic mathematical functions. Features of Matlab appropriate to digital
communication system modeling include random number generators, correlators, filter design functions, and FFT functions to estimate frequency content of random waveforms.

This simulation (and all digital communication simulations) operate at baseband so that the simulation sampling rates can be managed. Sampling at RF rates is simply not possible or even desirable.

All system parameter assumptions have been previously stated or are stated in the following sections. The model can be used to estimate baseband frequency content, and the effects on BER due to signal overlay or the effect on BER due additive white Gaussian noise specified as energy per bit / noise power (Eb/No). Eb/No is the SNR metric of choice as it is parameter easily related to specific system parameters such as data rate and carrier power. The additive white Gaussian noise disturbance is appropriate to modeling receiver thermal noise, and the noise induced by a spread spectrum adjacent channel IBOC signal.

The model is not currently coded to predict performance degradation due to multipath interference; such a channel model is comparatively complex and could be added with additional contract funding at a later date.

Specific baseband pulse filtering can be easily coded; the only filter available at this time is a square root raised cosine filter. Specific Gold codes used in a proposed system could be easily added if supplied; exact Gold codes used would be highly desirably if detailed performance results were desired in the future.

We verify (specific examples in other sections of this report where appropriate) the simulation code by comparing BER to Eb/No from the simulation to known perfect theoretical results from M-ary detection theory. The theory provides an absolute bound that can not be exceeded. The simulation will always be slightly degraded but close to theoretical results as the simulated signals will not be perfectly orthogonal and are finite in length. Any specific hardware simulation can be made to operate close to simulated results, although other losses may be incurred such as correlator insertion losses. For example, a proprietary system MDS possesses knowledge of was designed for a 15 dB processing gain, but measured about 11 dB in the hardware laboratory.

Comparison of simulation results to theoretical results is very powerful. No system can exceed theory; if the simulated system then operates at the edge of say some spectral mask, then a prototype system can do no better. This allows very broad statements to be made with confidence about any enhancements offered.

1. FM-1 NOMAC IBOC

In Figure 1. a comparison is shown for the theoretical biorthogonal (binary case) BER predictions (solid line), and the results from our baseband computer simulations (X's). We note a several dB loss of performance, which is expected as signal orthogonality is not perfect with finite length pseudo random sequences; the theoretical BER curve assumes infinite length signalling waveforms. Figure 1. is instructive and serves to validate our baseband computer simulation for NOMAC communication systems.

Several sources of performance degradation may be noted from our canonical transmitter/receiver system. First, the processing gain is 10 log10 (192 / 1) = approximately 23 dB, therefore the system will suffer from narrow band interference in its passband that exceeds about 23 dB, increasing the bit errors. Second, it requires an infinite length signal suite in order to achieve the perfectly uncorrelated white noise characteristic that the mathematical analysis assumes. A signal 192 samples in length, even if generated by a very good random number generator, possesses significant cross correlation structure, increasing the bit errors. Third, the waveform synchronization is critical; any mismatch between the signalling waveforms in time against the receiver stored waveforms in the correlator bank will degrade the error performance from theoretical predictions. Fourth, the idea of overlaying 48 noise waveforms is mathematically equivalent to a 48 user CDMA system with perfect power control. We can bound the performance using established CDMA multi-user access equations, then assess the degradation due to loss of orthogonality over a finite length noise signal via our computer simulation (the simulation includes the entire signalling suite of 48 waveforms. The text of [13] refers to a patent for perfectly orthogonalizing all 48 band pass filtered waveforms, however, that would only be true for perfect synchronization in time. Any synchronization errors would induce correlation amongst the signalling waveforms.

A more significant loss of performance is seen when 48 channels are summed in order to increase the aggregate channel data rate for transmission of stereo audio. In Figure 2., which plots the simulated single channel BER data (X's) against the simulated 48 channel BER (O's), we see that nearly 15 dB additional Eb/N0 is required for the same performance.

2. Enhanced IBOC System

We note that the enhanced system will suffer BER performance degradation as well, although not to the extent of a NOMAC system. Gold codes are nearly orthogonal, thus can approach theoretical limits of signal overlay. A more detailed multi-user analysis (ie, co-channel interference performance) is presented in the next section. We did, however, use an asymptotic mathematical expression, based on perfectly orthogonal codes, with noise power set to zero (equations in next section) to predict that the probability of bit error should be approximately .04 with 32 users, each with equal power. A long simulation run (32 users, nearly orthogonal random binary 64 bit codes, no added noise) estimated the Pe to be .07. This is in very close agreement, the difference due primarily to statistical variation and not quite orthogonal codes; such a simulation test serves to validate our FM-1 enhanced baseband simulation, providing extra confidence for use in spectral estimation, a more important issue in IBOC system analysis.

It is very important to note that perfect code synchronization is available in the baseband simulation to produce the results in the preceding paragraph. Indeed, through varying the simulation parameters, it was found as expected that performance decreased rapidly as increasing synchronization error (as a fraction of a chip) was inserted. This effect is easily seen from inspection of the autocorrelation function of a Gold code [13].

Based on Pe equations readily available [12], for M-ary biorthogonal signalling, the theoretical BER can be predicted and used as an upper bound on system performance. The Pe expressions for M-ary biorthogonal signalling are not closed form, awkward to evaluate, so we have reprinted a plot (Figure 6.) from [12] to summarize expected probabilities of bit error. These curves assume perfectly orthogonal signals, perfect synchronization, an AWGN channel, and rectangular baseband pulse shapes, and ideal matched filter receiver architecture. Obviously, real world performance will degrade in the presence of a mobile propagation channel, pulse shaping, and synchronization errors.

The performance loss in a real world fading mobile channel is significant. To present a simple comparison, we reprinted known theoretical results from [11] in Figure 7. Comparatively few theoretical results are available for fading channels, thus simulations are used to quantify performance predictions for specific modulation schemes, different signalling schemes, different baseband filtering methods, and varying mobile channel conditions, as all these factors affect the performance in highly nonlinear ways.

3. Baseband Spectrum Estimates

We used computer simulations at baseband in order to estimate the spectrum of the Gold code based enhanced FM-1 system. The NOMAC system was not analyzed for spectral content as its spectrum is controlled purely by the bandpass filtering (unknown filter parameters) of the random noise signalling waveforms, thus classical baseband pulse shaping can not be imposed.

The enhanced system that makes use of Gold codes represents a straightforward modification to our canonical simulation as it's simply a signalling waveform change; approximate Gold codes have been added as appropriate and simulation results used to estimate the baseband signalling spectrum, assuming raised cosine pulse shaping for spectrum control. Additionally, Monte Carlo simulations were performed to estimate performance degradation due to signal overlay self interference, with results consistent with asymptotic theoretical predictions.

The spectrum of a digitally modulated signal is governed by the data rate, baseband pulse shaping, the modulation scheme selected, and additional RF filtering applied in the transmitter (refer to Appendix for additional detail).

Any IBOC system will benefit greatly from lowering the data rate to rates below 100 kbps, obviously, and the available codecs are moving in that direction. Here, we assume a 96 kbps codec will be available for use in the near future. High quality compression of wideband audio (typically 7 kHz or 20 kHz bandwidth) will become increasingly important for digital commercial radio broadcasting and Integrated Services Digital Network (ISDN) applications.

128 kbps 20 kHz stereo is now available, and it APPARENTLY sounds quite reasonable, though not identical to a CD original. One product that offers this is the Comrex DX200. It is not known exactly how their codec works, but the audio coding approach that seems to be most popular these days might be called "suband-transform" coding. The transform would be something akin to a DCT, properly sized and overlapped to get good transient response, and the suband nature allows differential bit-allocation across the band in an attempt to place the quantization noise where it is least audible. It is our belief that MPEG now has standardized 3 layers of audio coding, and is working on a fourth. The sound system for digital television goes by the name of Dolby AC-3.

The bandwidth available to an IBOC system in a single sideband is 100 kHz; if a data rate of 192 kbps is desired (assume rate 1/2 FEC, and a date rate of 96 kbps), then the spectral efficiency is set at 2 bps/Hz. Digital modulators such as 16-ary PSK and 16-ary QAM can achieve 2 bps/Hz, and 32-ary PSK/QAM can achieve 2.5 bps/Hz. The penalty paid for M-ary PSK style system to achieve high bandwidth efficiency is reduced power efficiency. QAM is more power efficient than PSK, however, receiver complexity is increased. The FM-1 enhanced system, however, proposes to use M-ary biorthogonal signalling, with signal overlay, as discussed next.

For the FM-1 enhanced IBOC system, however, wherein biorthogonal signalling is used, vice PSK signalling, the bandwidth efficiency is achieved by (apparently) careful baseband pulse shaping, and signal overlay of 32 channels, each channel chip rate being 76.8 kcps (assumed) and a data rate per channel of 6 kbps (assumed). The signal overlay does not affect the power spectral density of the baseband signalling suite. M-ary orthogonal signalling or M-ary biorthogonal signalling increases power efficiency as M increases, but at the expense of bandwidth efficiency, thus the signal overlay procedure to recover "lost" bandwidth. We will find later in this report that baseband pulse shape filtering is critical to meeting spectral mask requirements, so that interference to adjacent channel IBOC signals and FM host main channel signals is minimized. By comparing the theoretical bandwidth efficiency of M-ary biorthogonal signalling to system bit rate requirements we will find the theoretical limit is exceeded, mandating use of a pulse shaping filter.

From Figure 3. we see that use of a beta=1 raised cosine filter [11] the baseband spectrum at 100 kHz is 50 dB down from the passband level. The parameter beta, or filter rolloff factor can be varied to achieve spectrum control, by trading off correlator losses and bit synchronization complexity if the bit stream is tightly filtered. For example, the IS-54 North American Digital Standard specifies a rolloff factor of .35 for a raised cosine pulse shaping filter.

3. Co-Channel Interference Analysis

In a wireless spread spectrum Code Division Multiple Access (CDMA) environment signals from a number of users on the same frequency channel arrive at the receiver input. A correlation receiver with a pseudo noise (PN) code matched to some desired user is used to separate one signal from the other signals; since the other users are assigned PN codes approximately orthogonal to all others the net effect is a background nearly white noise level. For this IBOC scenario of the NOMAC FM-1 system, the co-channel interference is actually self interference from the 48 channel signalling suite. In the FM-1 enhanced system based upon Gold codes the self interference consists of 32 channels of signalling. If the number of co-channel users is large enough so that the Central Limit Theorem may be invoked an expression may be obtained for BER that is very simple, especially if perfect power control is assumed (mutli-channel signalling as in this IBOC analysis is equivalent to perfect power control), wherein all signals arrive at the receive with equal power. For this report we will assume Gaussian statistics (number of users "large", say five or more at least), and consider both the equal power scenario and the unequal power scenario.

A widely accepted expression for Probability of Bit Error (Pe), that has held up in field trials is

Pe = Q{[3N/(P1/P0 + P2/P0 + ... +PK-1/P0 + N0/2TbP0)]1/2}

where Q(.) is the standard "Q-Function," N is the length of the PN code, Tb is the message bit interval, N0 is the noise power in Watts/Hz, P0 is the power in Watts of the desired signal, and Pk, k = 1, 2, 3, ... , K-1, is the power in the undesired signals representing the co-channel interference power. This expression assumes the interfering signals are of fixed number, K-1, and constant but unequal power. In a wireless scenario that is co-channel interference limited rather than noise limited, N0 can be set to zero for convenience; this is useful for producing quick estimates of performance degradation due solely to co-channel interference.

A very simple expression obtains by setting noise power to zero and assuming all signals arrive at the receiver with equal power Pk = P0 Watts. This case results in

$Pe = Q[(3N/(K-1))1/2]$

from which it is clear that either a 48 channel or 32 channel signal overlay will experience self interference. For example, for K = 48 signals, N = 192 PN code length, we find Pe is approximately .0006; adding noise, N0, to this example would decrease the value of the argument in Q(.), increasing Pe measurably, as would be expected. The enhanced FM-1 system, consisting of 32 signals without additional spectrum spreading would be expected to experience a Pe = .04 using a 32 bit code for computational savings. Our baseband simulation estimated the BER at approximately .07, so we are quite confident in our simulation accuracy. We can easily insert any desired code length in the baseband simulation or code construction desired.

5. Adjacent Channel and Host Interference Issues

The IBOC signal must be positioned in the FM sidebands beneath a spectral mask as determined by the FCC so as to not interfere with either the host FM signal or adjacent channel signals. The appropriate spectral mask will not be repeated here as it is commonly available in other reports (47 CFR 73.317, Appendix B), but basically requires the digital sideband to reside within 120 kHz to 240 kHz from the main channel (host) unmodulated carrier with spectral magnitude at least 25 dB below the host carrier. Note that this mask refers to spurious emissions; digital sideband spectra that meet the minimum criteria of this specification may interference with adjacent channel IBOC stations and/or host FM main channel audio, thus a tighter bandwidth may well be required in practice for an IBOC signal. A 32 channel, 32-ary 192 kbps digitally modulated signal can not be viewed as a spurious emission.

Other potential problems resulting from the digital sideband signal include interfering with a third harmonic of the 38 kHz subcarrier (receiver manufacturer specific) and interfering with a potential 92 kHz SCA.

The enhanced FM-1 IBOC spectrum is governed by the baseband pulse filtering, the per channel bit rate, the 32-ary biorthogonal modulation choice, and any subsequent RF filtering. The power (area beneath the PSD) is obviously a function of the bandwidth, spectral shape, and peak transmitted power in the digital sideband.

As can be seen from Figure 3., the baseband signalling spectrum can be tightly controlled by careful pulse filtering, and selection of bit rates. The spectrum measured by and reported in [14] indicates clearly that spectral energy is visible below the 120 kHz lower limit at some level difficult to accurately gauge, and exceeds the upper 240 kHz limit by some margin, even exhibiting energy beyond 250 kHz. This spectral energy would adversely affect an adjacent channel IBOC signal. It is not known by this author, with respect to the spectral plot viewed from [14], what baseband pulse filtering was used (if any), what precise bit rates were used, and what RF filtering was imposed. The spectrum can be modified significantly by varying such design parameters; the baseband spectrum estimate shown in Figure 3., indicates that use of a raised cosine pulse filter, combined with a per channel chip rate of 75.6 kcps (assuming 32-ary biorthogonal modulation and 64 bit Gold codes), may bring the digital sideband spectrum within the mask boundaries.

The baseband pulse filtering can be critical to spectrum control. In our simulation, we created a 32 signal overlay of 64 bit codes, both filtered (beta = 1 raised cosine filtering), and unfiltered. The spectrum shown in Figure 4. is unfiltered, and the excessive spectral energy outside the mask limits is clearly visible. By comparison, the spectrum shown in Figure 5., estimated from a 32 signal overlay of raised cosine filtered 64 codes is very well behaved, falling well with the 120 kHz to 240 kHz mask limits, with greater 50 dB attenuation from passband at the mask boundaries. As a reminder these spectrum were produced by a baseband simulation with 76.8 kcps per channel chip rate. The baseband signalling suite is subsequently modulated to the sidebands, thus close attention must be paid to any potential bandwidth expansion from the modulator. We assume the baseband spectra is translated to the sidebands used for IBOC transmission without bandwidth expansion by use of DSB modulation. This modulation scheme would replicate the digital signal redundantly into a lower and upper sideband, allowing use of the diversity idea proposed in [13] in order to improve digital audio performance.

Indeed, the theoretical limit (assuming no pulse filtering) is somewhat exceeded by the M-ary biorthogonal signalling system proposed, thus pulse filtering is absolutely required to meet spectral mask constraints. From [12], the theoretical maximum bandwidth efficiency for M-ary biorthogonal signalling using rectangular pulses is 4Log2M/M bps/Hz, which for 32-ary signalling is 5/8 bps/Hz. A 32 signal overlay plus spectrum spreading of 64/5 results in a maximum bandwidth efficiency of (5/8)(32)(5/64) = 25/16 bps/Hz, slightly less than the 2 bps/Hz required for this proposed IBOC scheme to fit safely within the spectral mask, thus not causing interference to adjacent channel IBOC systems or the host FM station main channel. Note that 25/16 bps/Hz is a theoretical maximum bandwidth efficiency for 32-ary biorthogonal signalling, and the real world efficiency will be lower. From the preceding analysis we see that baseband pulse shaping is mandatory.

The power in the digital sideband(s) can be set to bring the spectrum beneath the mask upper limit and/or control interference to subcarriers to limit the PLL decoder problems. Obviously, lowering the power lowers the per bit Eb/N0, with a resultant increase in BER.

The IBOC FM-1 enhanced system design as postulated in this report, based upon bit rates and signalling specifics detailed in the system description section of this report, would exhibit a different digital sideband spectrum than apparent in [14]. For example the pulse shaping filter parameter(s) can be varied to control baseband spectra, the RF filtering can be used to control transmitted signal spectra, the FEC rate can be used to lower the aggregate channel rate, the length of the Gold code can be used to control the per channel chip rate to advantage, and the M-ary signalling can be used to BER against baseband bandwidth. These design factors are highly interrelated of course, and any measure used to reduce digital sideband bandwidth will inevitably increase the BER, potentially below accepted performance expectations. Ultimately, however, the promise of improved codecs (ie, lower bit rate), will buy performance trades in favor of any IBOC system in use.

6. Narrow Band Interference Analysis

One expects that the IBOC receiver would not experience significant interference from NB signals, rather most interference should result from spectral leakage from the host FM station, and adjacent channel FM stations with IBOC capability. In the event NB interference may be an issue, a brief summary of the appropriate equations is given.

The output SNR in dB, (SNR)o, for a generic correlation DS receiver, may be related to Processing Gain (PG) in dB, and input SNR in dB, (SNR)i, by the expression

$$(SNR)i = (SNR)o - PG$$

where Processing Gain is defined as the ratio of spread spectrum signal bandwidth to message signal bandwidth. If we assume rectangular pulses for message bits, I/Q modulation, for the basic FM-1 NOMAC system, the processing gain in dB is given by

$$PG = 10 \mathrm{Log}(192) = 23 \text{ dB approximately,}$$

although the true PG is usually lower than the theoretical due to a residual carrier component from the modulator. The enhanced FM-1 system, based upon a set of 32 waveforms consisting of 32-ary 64 bit (assumed) coding, is a spread spectrum system with a low processing gain of 10Log(76.8/6) = 11 dB, thus NB interference rejection is limited to 11 dB.

The output SNR of a correlator receiver is given by

$$(SNR)o = P/[(N0/2Tb) + (J/PG)]$$

where P is the power in Watts at the receiver input due to the desired transmitted DS signal, N0 is the noise power in Watts/Hz, Tb is the message bit interval, PG is the processing as previously defined, and J is the power in Watts due to a NB interference source.

Substituting Eb = PTb, and rearranging terms slightly, we can write

$$Eb/N0 = [1/2 + J/(N0B)](SNR)o$$

where B is the total DS signal bandwidth in Hz, in order to directly relate BER to NB interference power in Watts and noise power in the receiver bandwidth. We see from this expression that if the interference power, J, is narrow band then increasing the spreading (ie, increasing B), improves performance. Note, however, that if the interference power is wide band then increasing the spreading does not improve performance because the interference power increases with the bandwidth.

7. Digital Receiver Complexity

Functionally, the receiver must accomplish several tasks generic to most digital data communication systems, although some processing tasks may not always be required. The receiver may be required to extract and track doppler information, and must always demodulate the carrier (at IF) in order to produce a soft bit stream suitable for further processing; obviously, the soft bit stream will require further processing before hard bit decision are declared. For a digital signal processing based receiver, the digital data demodulation may be performed in software or dedicated digital hardware.

The retrieval of data symbols from the received signal will involve demodulation of data modulated via binary phase shift keying (BPSK), quadrature phase shift keying (QPSK), offset QPSK (OQPSK), minimum shift keying (MSK), continuous phase modulation (CPM), Gaussian minimum shift keying (GMSK), frequency shift keying (FSK), quadrature amplitude modulation (QAM), amplitude shift keying (ASK), or a wide variety of other related modulation schemes in use today for modern digital communications systems [1,10,11,12]. As bandwidth efficiency, and power efficiency, often competing design constraints, are of paramount importance for digital data communication systems, ever increasing digital signal processing based receiver complexity is a certainty.

Data demodulation is a necessary step in extracting digital data from the received signal, but many other processing steps are usually invoked to improve receiver performance, especially in harsh multipath propagation communications environments. For example, so called RAKE receivers [9] take advantage of the information contained in multipath components to improve bit error rate performance at the price of processing complexity. Almost all time division multiple access (TDMA) systems require channel equalization to mitigate the effects of intersymbol interference [8]. Code division multiple access (CDMA) systems may not require sophisticated channel equalization, but do require code correlation processing. Literally hundreds of research papers and technical books are available on these advanced topics. Since the multipath communications channel is a time-varying system, adaptive digital filter processing is required for best performance, further increasing the complexity of modern digital data communication systems. If additionally, the receiver is based upon digital signal processing methods, the complexity is increased more so as the algorithms must adapt and execute in real-time.

A significant complexity issue for the USADR proposed enhanced system is the requirement for 512 correlators, each 32/64/128 bits in length, depending on the specific design. Correlators, perhaps a SAW device, are comparatively expensive with significant power consumption and insertion losses . In this case, 512 correlators is an impractically high number of separate correlators, so these would have to be implemented in some sort of very high performance digital processing system.

7. RAKE Receiver Complexity

In spread spectrum CDMA systems, due to the relatively high bandwidth of the signals (of course with increased complexity) several different propagation paths can be resolved. In such cases the RAKE receiver [9] can be put to good use in order to decrease the bit error rates. The RAKE receiver in its simplest form is a linear weighted sum with time-varying coefficients that are a function of the attenuation and delay of the individual multipath components. Since multipath propagation is highly time dependent, especially in a mobile communications environment, the coefficients must be estimated from an estimated channel impulse response and updated in real-time to accurately model the current channel propagation characteristics.

The coefficients are a function of a linear convolution that can be implemented via standard FFT based high speed convolution algorithms with well defined computation complexity. The complexity trade-offs will be a function of data block size (ie, FFT length) and how often the coefficients require update, which is a function of channel dynamics; computer simulation must be utilized to model the specific multipath communications scenario and define data block sizes and update rates. The preceding operations lend themselves to conventional ASIC implementation that can handle approximately 100 million multiply/add operations per second.

A specific example of a RAKE receiver DSP design is given in [3], for a spread spectrum communication systems application in a multipath environment. A comparatively high chip rate is used (approximately 20 Mchips/second) in order to resolve the relatively small delay spread of the indoor channel in the specific application considered. For this application a data rate of 16 kb/s is invoked, the digital modulation chosen is BPSK, and Gold codes are used for spreading, a common choice; the receiver is capable of resolving 8 multipath propagation paths, and the correlator operates digitally at baseband in time in parallel over each of the 8 RAKE 'arms.' The receiver is based upon two time integrating correlators providing 8 parallel channels into a TMS320C25 executing with a 100 nsec cycle time.

Figure Titles

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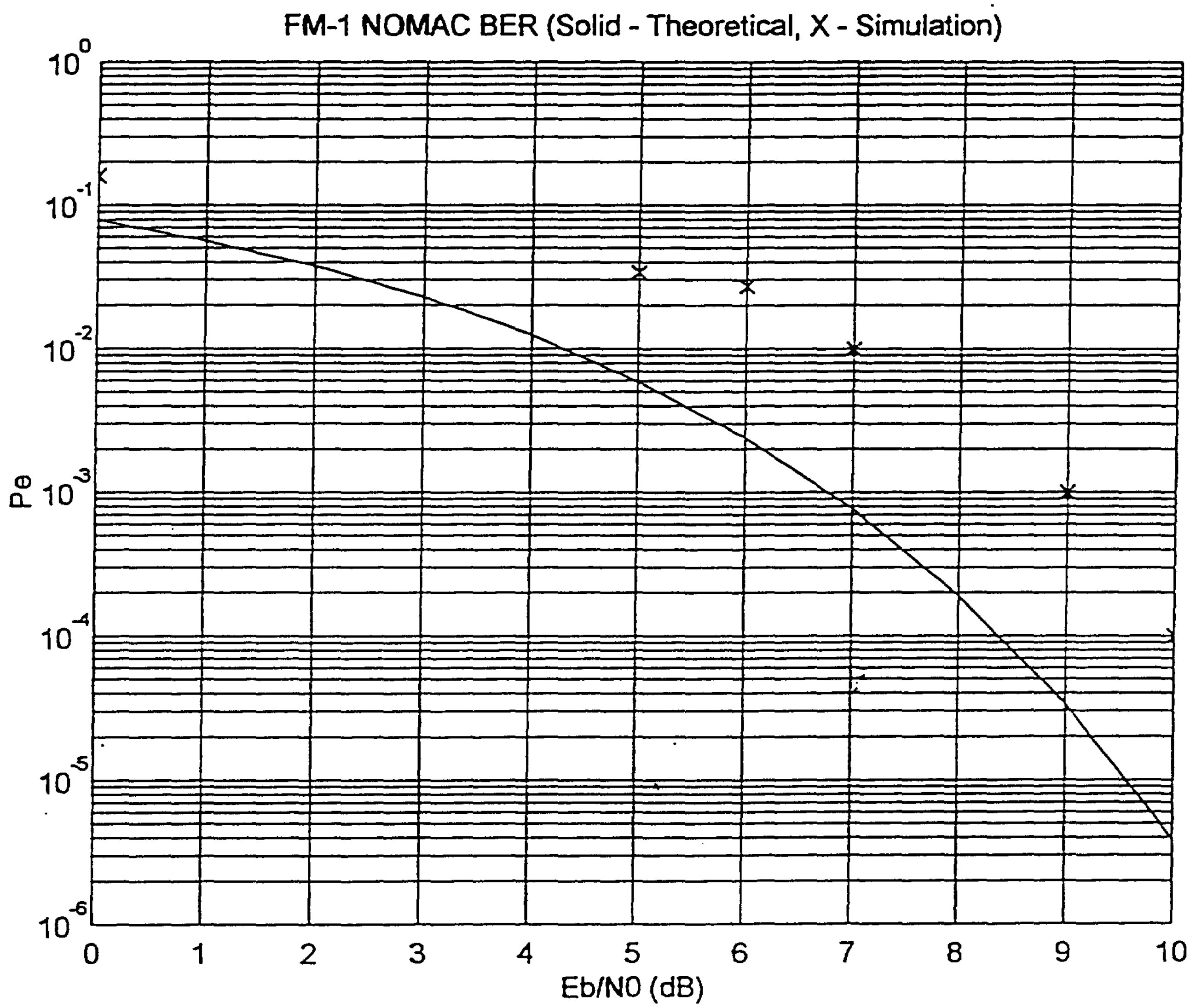

Figure 1. Simulated BER For FM-1 NOMAC System Vs. Theoretical BER For Binary Biorthogonal Signalling.

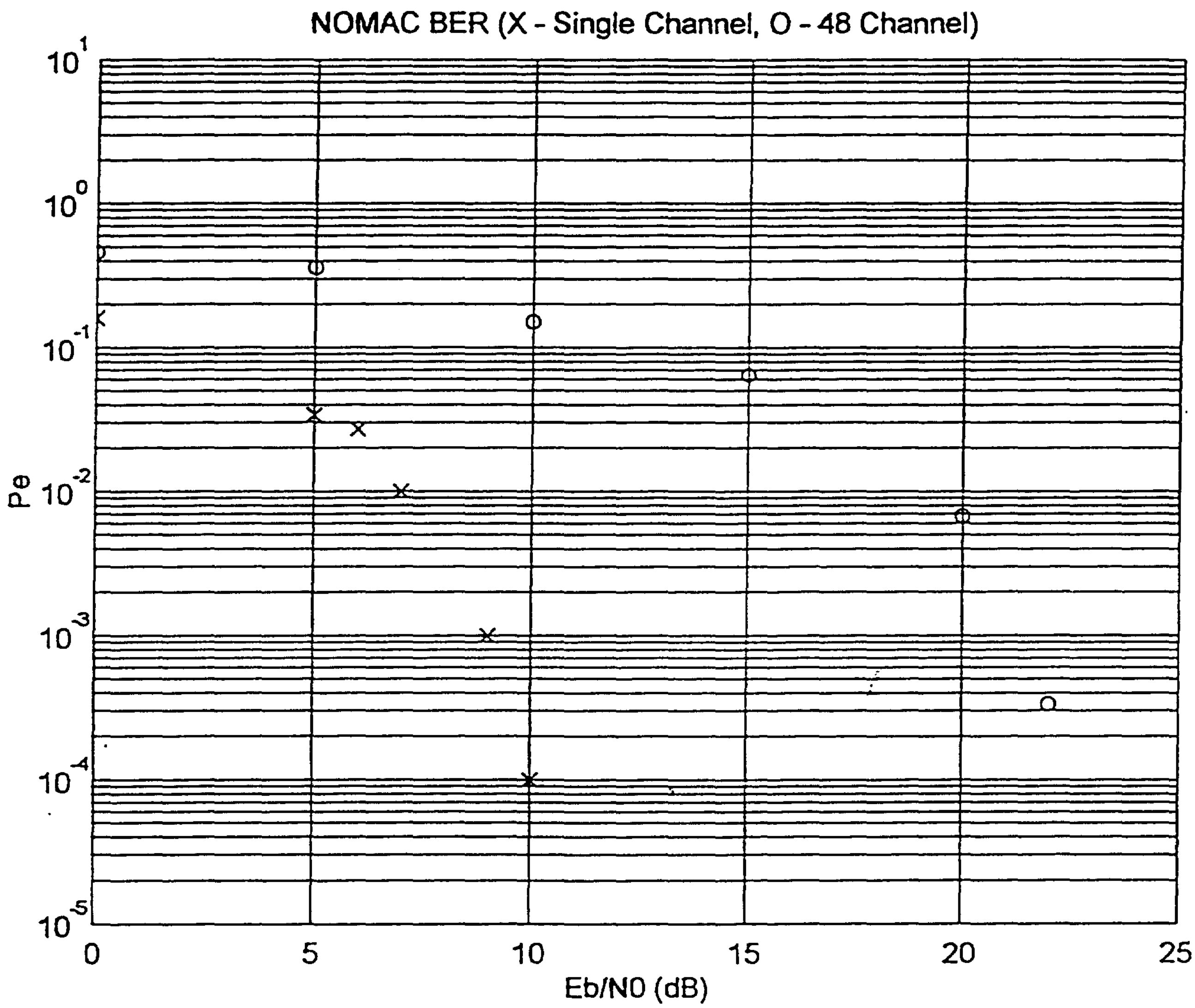

Figure 2. Simulated BER For FM-1 NOMAC Single Channel Performance Vs. 48 Channel Signal Overlay.

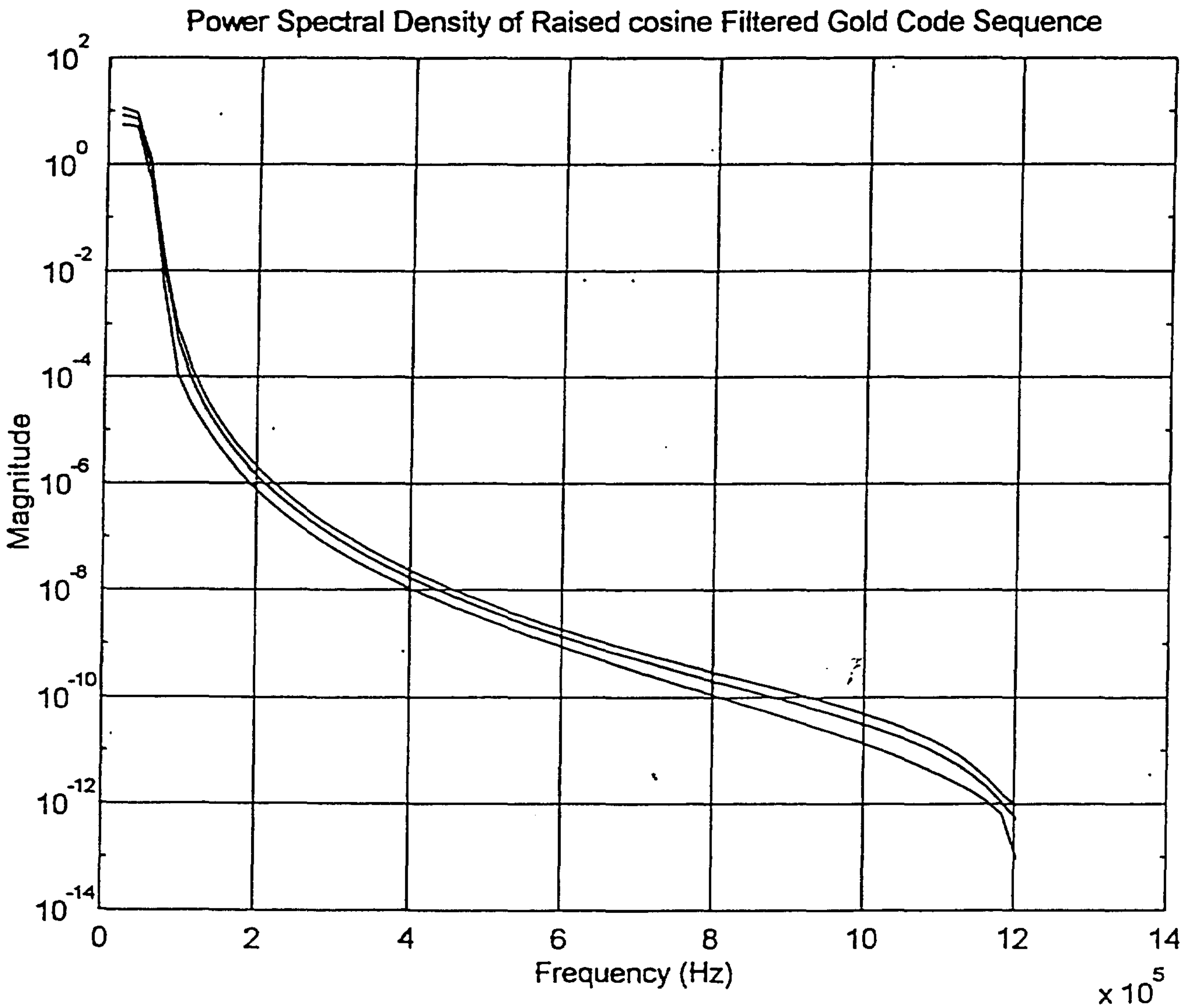

Figure 3. Baseband Spectrum Of NRZ Gold Code Sequence With Beta = 1.0 Raised Cosine Filtering.

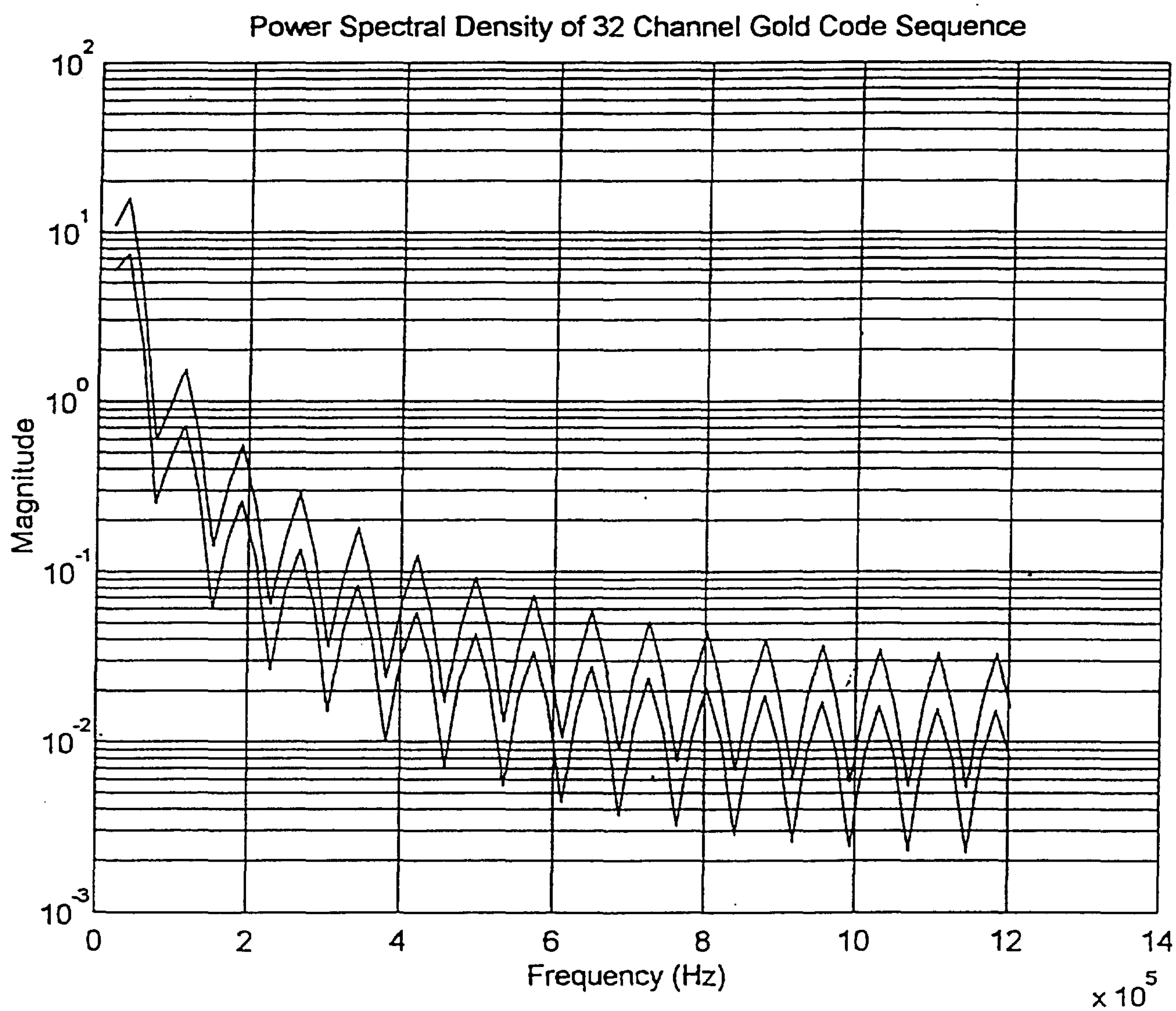

Figure 4. Baseband Spectrum Of NRZ Gold Code Sequence With Overlay Of 32 Signals, No Filtering.

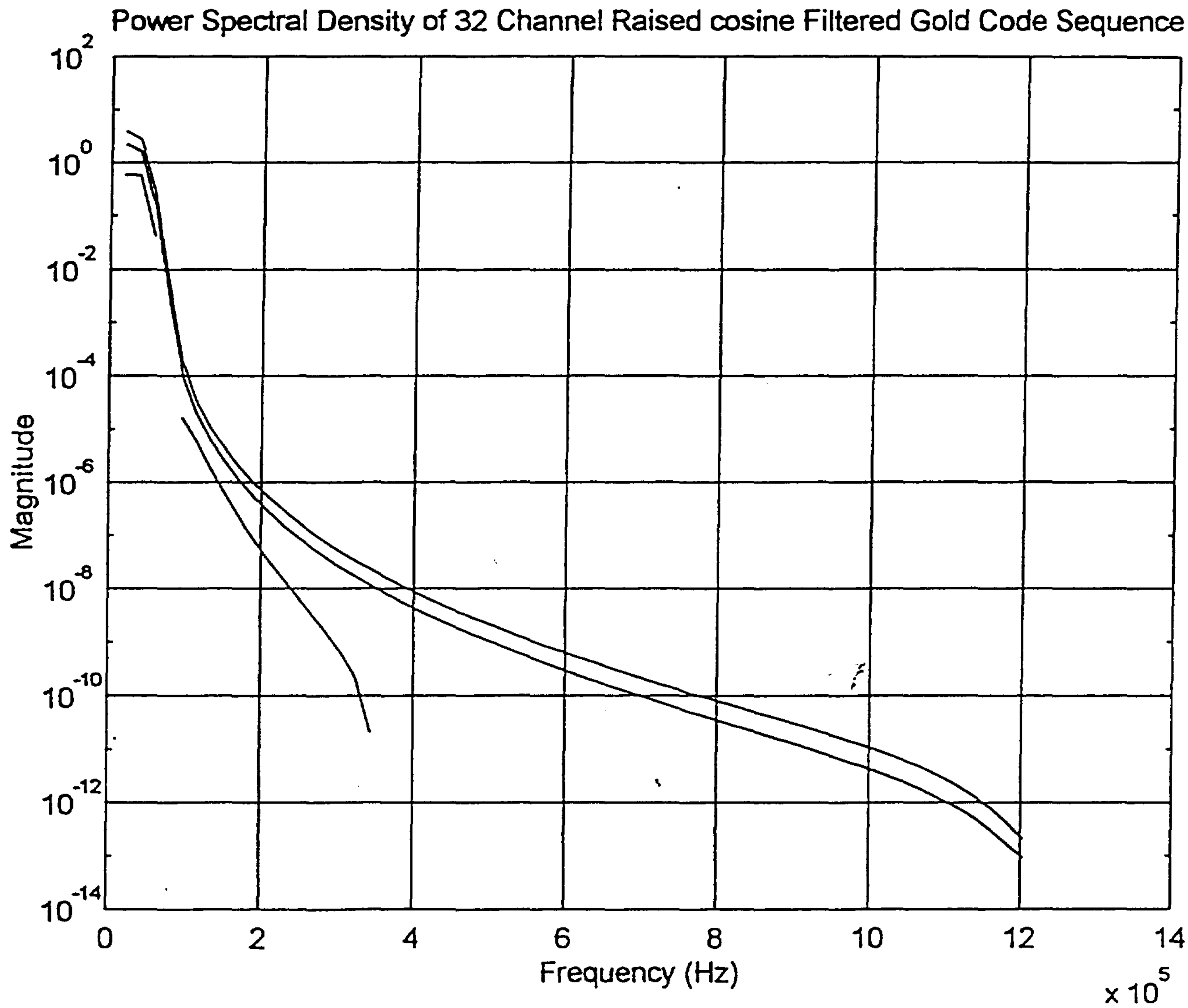

Figure 5. Baseband Spectrum Of NRZ Gold Code Sequence With 32 Signal Overlay, Beta = 1.0 Raised Cosine Filtering.

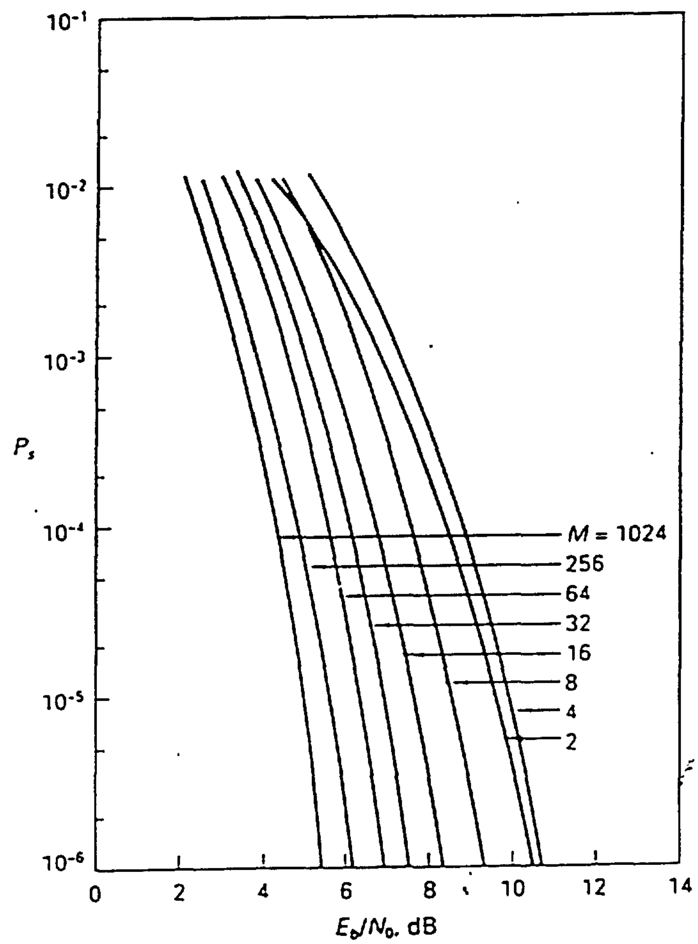

Figure 6. Theoretical BER Curves For M-ary Biorthogonal Signalling (Reprinted From [12]).

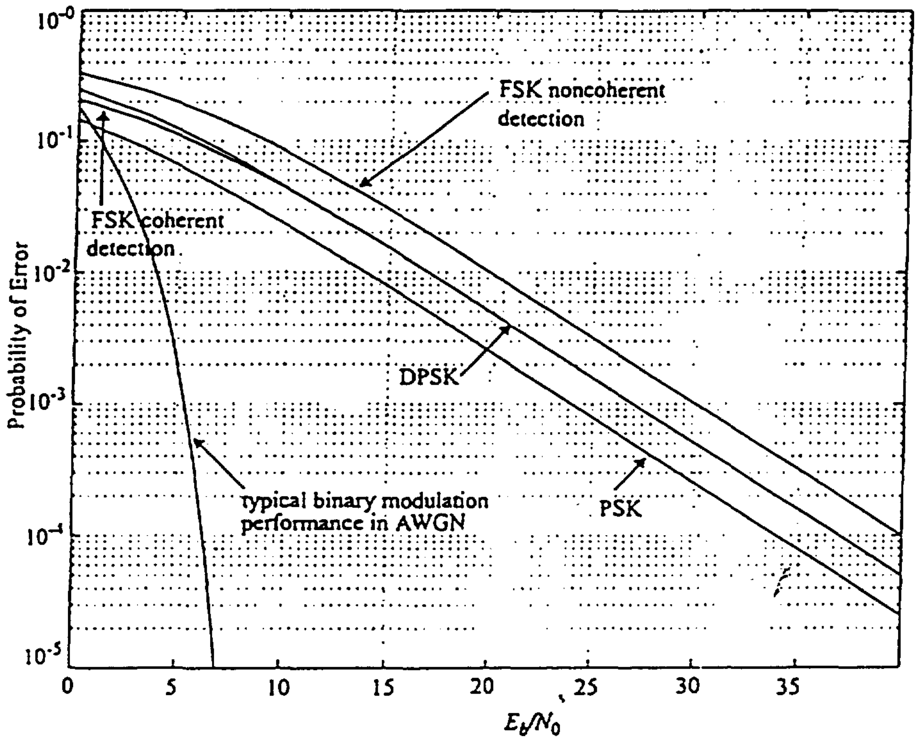

Figure 7. Theoretical BER Curves Comparing Performance In AWGN Vs. A Fading Channel (Reprinted from [11]).

APPENDIX 12
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ROBUST IBOC DAB AM AND FM TECHNOLOGY FOR DIGITAL AUDIO BROADCASTING

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ABSTRACT

A robust In-Band On-Channel (IBOC) Digital Audio Broadcast (DAB) System for improved performance over existing AM and FM broadcasting is under development by Westinghouse for USA Digital Radio. The solution is both forward and backward compatible without the allocation of additional channel spectrum. Broadcasters can simultaneously transmit both analog and digital signals within the allocated channel mask allowing full compatibility with existing analog receivers. The solution also allows broadcasters to transmit an all-digital signal, replacing the hybrid analog/digital signal. The solution is tolerant of interference from adjacent channels, or interference from the co-channel analog transmission, even in a multiple station, strong-signal urban market. This paper describes spectral occupancy, power ratios, modulation formats and coding as well as the introduction of frequency and time diversity. It also addresses the adoption of a forward compatible all-digital transmission for both AM and FM broadcasting.

I. INTRODUCTION AND BACKGROUND

Digital Audio Broadcasting is a medium for providing digital-quality audio, superior to existing analog broadcasting formats. The advantages of digital transmission for audio include better signal quality with less noise and wider dynamic range than with existing FM and AM radio. The goal of FM DAB is to provide virtual-CD quality stereo audio along with an ancillary data channel with optional capacity up to 64 kbps depending upon a particular station's interference environment. The goal of AM DAB is to provide stereo audio with quality comparable to present analog FM quality and a 2.4 kbps ancillary data channel. The development of new high-quality stereo codec algorithms indicates that virtual-CD stereo quality will soon be practical at rates as low as 96 kbps while stereo audio, startlingly superior in quality to existing AM audio, can be attained at 48 kbps. IBOC requires no new spectral allocations because each DAB signal is simultaneously transmitted within the same spectral mask of an existing allocation. IBOC DAB is designed, through power level and spectral occupancy, to be transparent to the analog radio listener. IBOC promotes economy of spectrum while enabling broadcasters to supply digital quality audio to their present base of listeners.

An independent technical evaluation conducted by *the Deskin Research Group* in 1996 revealed various weaknesses [1] in the previously proposed FM IBOC systems [5]. These deficiencies included DAB interference to host, first and second adjacent interference, and lack of robustness in multipath fading. These deficiencies were addressed in subsequent development work as reported in [2], and substantially eliminated in the new design which has evolved over the past year since the *Deskin* study.

The design of the AM DAB is continuing as planned at *Xetron*. A very brief overview of the AM IBOC DAB system is also presented in this paper.

II. FM OFDM IBOC SYSTEM DESCRIPTION

A brief description of the IBOC simulation model is presented here. Several modulation techniques were evaluated for the IBOC DAB application, including multicarrier spread spectrum, high-rate single carriers and Orthogonal Frequency Division Multiplexing (OFDM). Tradeoff analyses led to the selection of OFDM. OFDM modulation has been shown to be tolerant of multipath fading when used in conjunction with FEC coding and interleaving. Furthermore, OFDM can be tailored to fit an interference environment that is nonuniform across frequency while also providing flexibility for additional optional subcarriers.

The DAB signal is transmitted on OFDM subcarriers located on either side of the analog spectrum. A spectral mask along with the FM and DAB power spectral densities is presented in Figure 1. Note that the FM spectral mask is defined as peak power measured in a 1 kHz bandwidth over any 5 minute interval. The power spectral density for the FM signal was empirically determined by power-averaging the FM spectrum over 5 minutes. Five stations in the Baltimore/Washington area exhibited the triangular power spectral density with a slope between 0.35 and 0.38 dB/kHz. Interestingly the stations measured included diverse signals ranging from "heavy metal" music to talk. The average slope of the 5 stations is 0.36 dB/kHz, which is assumed for the modulated spectral plots.

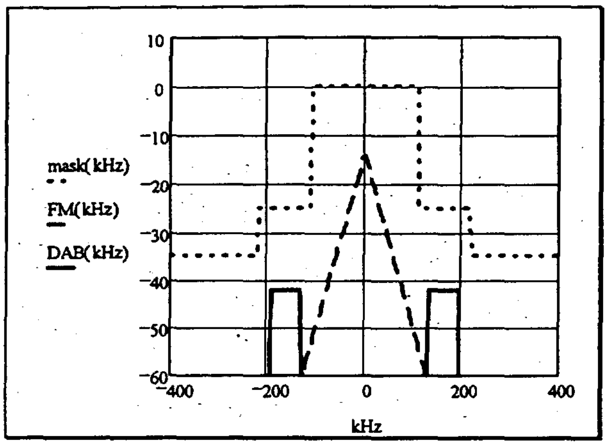

Figure 1. Power spectral densities of FM and DAB signals below FM spectral mask.

The total FM power can be found by integrating the triangular power spectral density.

$$P_{total} = \int_{-\infty}^{\infty} P_{peak} \cdot 10^{-0.36|f|/10} \cdot df = 24.12747 \bullet P_{peak}$$

Then the peak of the FM power spectral density is located 13.8 dB ($10 \cdot \log(24.12747)$) below the total carrier power reference level (0 dB) as shown in Figure 1. The DAB power level on each side of the FM spectrum is placed 25 dB below the total FM power (this value is adjustable by the broadcaster to accommodate special interference situations). The DAB density in a 1 kHz bandwidth can be calculated. The power spectral density of the DAB signal can be very closely approximated by dividing its total power by its effective Nyquist Bandwidth.

$$PSD_{DAB} = \frac{10^{-25/10}}{81 \cdot 0.796875} = 4.9 \cdot 10^{-5}$$

Then the power spectral density of the DAB signal in dB as shown in Figure 1 is computed to be -43 dB/kHz ($10 \cdot \log(4.9 \cdot 10^{-5})$).

The baseline DAB system assumes 81 subcarriers above and 81 below the host FM spectrum. Each DAB subcarrier is QPSK modulated at a symbol rate of 750 Hz. The inphase and quadrature pulse shapes are root raised cosine tapered (excess time=17/16) at the edges to suppress the spectral sidelobes. Although this pulse shape reduces the throughput capacity relative to the rectangular pulse by 5.88%, performance in multipath is improved and the resulting spectral sidelobes are reduced, lowering interference. This pulse shape results in orthogonal subcarrier frequency spacing of 796.875 Hz. A plot of the pulse shape normalized to 1 unit of time is presented in Figure 2.

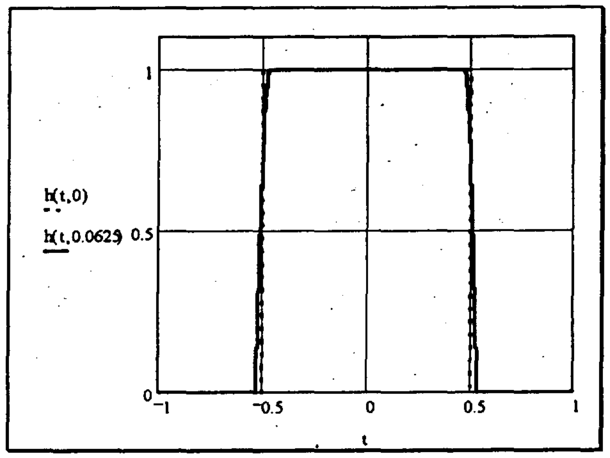

Figure 2. Plot showing rectangular Nyquist pulse (dotted) and the root raised cosine tapered pulse (solid).

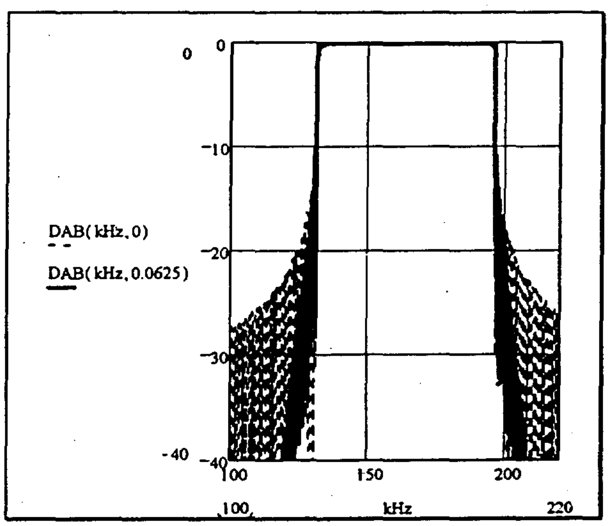

Figure 3. Improved spectral sidelobe suppression of Nyquist root raised cosine tapered pulse (solid) and rectangular Nyquist pulse (dotted).

Figure 3 shows plots of the DAB spectra using both the rectangular and root raised cosine pulse shapes. . The 81 subcarriers in this case are Nyquist-spaced at 796.875 Hz The effective Nyquist pulse time is approximately 1.2549 milliseconds for both cases yielding an effective Nyquist bandwidth of 796.875 Hz.

Potential subcarrier locations are identified by their offset from the host FM center frequency. For example, a reference subcarrier is placed at 144234.375 Hz above and below the FM center frequency. This reference subcarrier position is number 181 from the FM center frequency (181*796.875=144234.375 Hz). The reference subcarrier is used to aid in acquisition and tracking of the symbol timing and carrier frequency. A narrowband phase locked loops tracks the reference subcarrier, effectively rendering it immune to dynamic fading conditions. The reference subcarrier also provides the local reference to initiate differential decoding at the receiver.

Subcarriers 182 through 245 carry 96 kbps. Subcarriers 165 through 180 can carrier an additional 24 kbps of FEC coded bits to create an effective code rate of R=4/5 on each side of the FM signal. The placement of DAB at ± 15 kHz about 114 kHz is avoided in the baseline system in order to reduce the noise introduced into inadequately filtered receivers. However the broadcaster will have the option to utilize this portion of the spectrum to improve robustness of the digital audio signal and/or to provide additional datacasting capacity.

Although each DAB sideband can be demodulated independently of the other, the intention is to combine the two sidebands, yielding a power gain of 3 dB, plus additional coding gain achieved by a R=1/2 code over the 4/5 coding gain on each independent sideband.

The total capacity of each sideband is 120 kbps (uncoded). After R=4/5 rate FEC coding, the coded capacity is 96 kbps for each redundant sideband. This data rate is sufficient for transmission of virtual-CD quality music plus a modest datacasting capacity. Optionally, additional carriers can be added to increase the datacasting capacity. These carriers would be located closer to the host analog FM signal.

III. INTERFERENCE ANALYSIS

The interference to and from the first adjacent channels placed ± 200 kHz from the host signal can be derived from the relationship of the adjacent signals shown in the plot of Figure 4. FM stations are geographically placed such that the nominal received power of an undesired adjacent channel is at least 6 dB below the desired station's power at the edge of its coverage area. Then the D/U (desired to undesired power ratio in dB) is at least 6 dB. Knowledge of the ratio of each station's DAB signal power to its FM host permits assessment of first adjacent interference to DAB. Similarly the interference of the first adjacent DAB to the host FM signal can be assessed from the relationship.

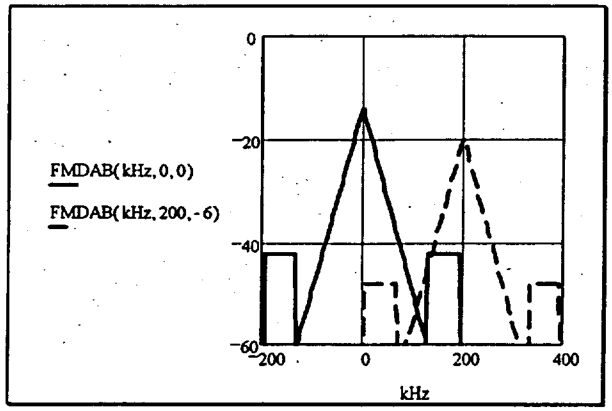

Figure 4. Interference scenario showing first adjacent at -6 dB (worst case edge of coverage).

Figure 5 illustrates the need for DAB spectral sidelobe suppression and bandlimiting due to the second adjacent DAB interference to the host DAB signal. At a station's edge of coverage, a second adjacent's nominal power can be up to 20 dB greater than the host's nominal power.

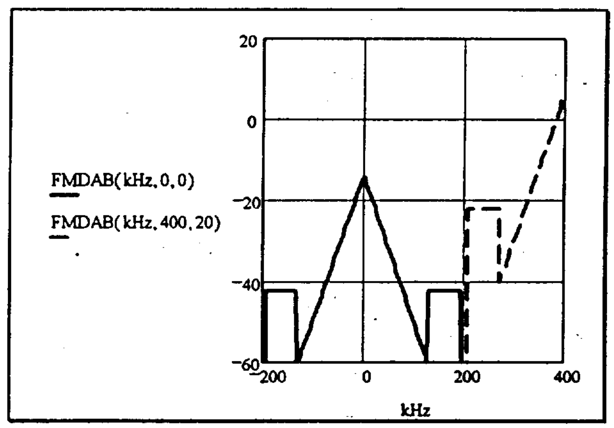

Figure 5. Interference scenario with second adjacent at +20 dB.

The effects of the various interference scenarios illustrated here are quantified through analysis and supported through simulation and testing. Analysis of the DAB to first adjacent interference at the edge of coverage showed that the total DAB signal should be set about -22 dB relative to its FM power.

The solution to the first adjacent interference problem is to place redundant, although not identical, DAB signals on either side of the carrier. Although the potential capacity is halved with this redundancy, interference problems are substantially reduced and substantial coding gain is achieved after combining both halves. A survey of existing U.S. radio allocations shows that it is very unlikely that both upper and lower adjacent channel interferers are present at their maximum interference levels (-6 dB) at the same geographic location within the host's coverage area. This frequency diversity is especially useful when multipath interference or spectral notches affect one sideband or the other.

A variety of simulations and analyses have characterized performance of the host FM signal in the presence of IBOC DAB. Specifically, main audio channel performance, SCAs, adjacent channels, and stereo subcarrier demodulation were investigated with an IBOC DAB signal appended to the host FM.

Main audio channel performance

Simulations have provided valuable insight into the character of FM post-detection noise in the presence of an IBOC DAB signal. For instance, results indicate that the audio noise level increases with the deviation of the FM signal. In fact, Figure 6 illustrates a significant rise in the post-detection noise power spectral density (PSD) as the FM deviation varies from minimum to maximum in the presence of an IBOC DAB signal placed between 78 kHz and 197 kHz from the FM carrier.

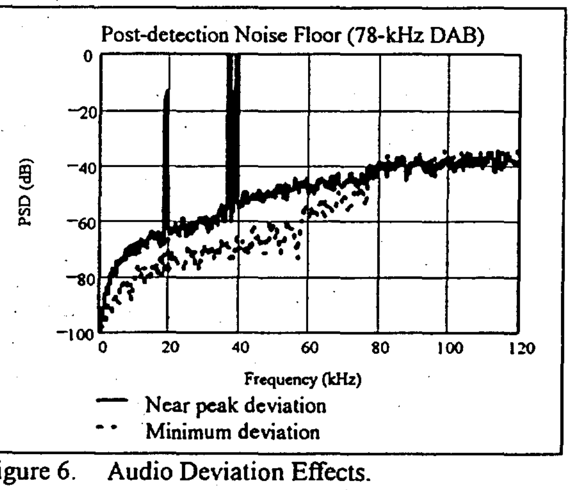

Figure 6. Audio Deviation Effects.

The nonlinear FM detector is responsible for intermodulating overlapping portions of the host FM and DAB spectra. The products are folding back into the post-detection audio band and raising its noise floor. Similar observations and conclusions were independently reached by the Electronic Industries Association (EIA) during their IBOC DAB testing [4].

Although these results are intriguing, they do not predict a degradation in host FM audio quality due to IBOC DAB. Because the DAB-induced post-detection noise floor increases in proportion to the deviation of the FM signal, the effect is self-masking: audio noise will be lowest during quiet passages, and highest only when the audio is loudest. Simulations have demonstrated this phenomenon.

The absolute level of host FM degradation will depend on the particular configuration of DAB. To determine the relationship between DAB location and audio signal-to-noise ratio (SNR), a number of performance tests were run when DAB noise would be most audible – during quiet passages of minimum FM deviation. Simulations were performed in which the receiver audio dynamic range was measured with only a 10%-deviated, 19-kHz-pilot-modulated FM signal and a DAB signal input to an FM stereo receiver located at the transmitter. The total power of the DAB signal was 22 dB below the power of the FM carrier. In the first four tests, the DAB was modulated using orthogonal frequency-division multiplexing (OFDM) with 4750-symbol-per-second quadrature phase-shift keying (QPSK) subcarriers using rectangular pulse shaping. The fifth test employed DAB with four times the number of OFDM carriers -- each occupying one-fourth the bandwidth (1187.5 Hz) -- and root-raised-cosine pulse shaping (to reduce spectral sidelobes that interfere with the host FM). In each test, the spectral occupancy of the DAB signal was changed: the start frequency was varied with respect to the FM center frequency, while the stop frequency was fixed at 197 kHz. Table 1 summarizes the results.

Table 1 - Audio Dynamic Range at Transmitter (peak-to-noise-floor SNR)

DAB start frequency	Audio SNR (dB/15 kHz)
78 kHz	64.7
100 kHz	67.3
124 kHz	68.3
129 kHz	68.8
129 kHz, pulse shaped	77.6

These results indicate that moving the DAB away from the FM carrier, increasing the number of DAB carriers, and pulse shaping the transmitted DAB symbols to reduce spectral sidelobes will significantly improve the performance of the host FM. Modulation and coding characteristics of the DAB signal can be traded for spectral occupancy to meet these goals.

Note that the new DAB baseline employs subcarriers spaced at 796 Hz which would improve performance over the carrier spacings reported here.

Audio simulations have verified that an SNR of 77.6 dB during quiet passages should render DAB-induced audio noise imperceptible to the listener. Furthermore, implementation constraints limit the SNR of typical receivers to around 60 dB. The noise engendered by these receivers will mask any degradation caused by DAB. The -22-dB, 129-kHz pulse-shaped DAB configuration is used as the baseline for the balance of this discussion.

SCA performance

SCAs (Subsidiary Communications Authorization) are optional channels multiplexed onto the baseband stereo spectrum from 53 kHz to 100 kHz. The SCA signal, which can be analog or digital, is transmitted by some FM stations for the use of private subscribers who typically pay for program material. Simulations were used to determine the impact of SCAs on IBOC DAB host FM performance, and to determine the impact of DAB on the performance of SCAs. SCAs with 10% deviation at 67 kHz and 92 kHz were simulated because they represent a large percentage of operational subcarriers.

In the current analog FM system, SCAs generally cause negligible interference to the host FM signal. However, when DAB is present, the addition of SCAs could increase the host FM audio noise floor due to the DAB/FM intermodulation effect described above. Figure 7 illustrates stereo subcarrier sensitivity to 92-kHz SCAs when subject to a pulse-shaped (PS) DAB signal starting at 129 kHz. In this case, the 92-kHz SCA reduces the host FM audio SNR from 77.6 to 69.8 dB; however, this noise level is still too low to produce audible effects. Figure 8 shows that SCAs located at 67 kHz have even less impact on audio performance.

Due to their location at the high end of the baseband spectrum, some SCAs currently operate at low SNRs because the post-detection noise floor increases with the square of the frequency. When DAB is added, the deviation of a wideband host FM signal into its IBOC DAB signal produces intermodulation which increases the post-detection noise floor, particularly in the higher baseband frequencies (since this is nearest the location of the pre-detection DAB). Moreover, the noise masking effect described above does not apply for SCAs, since their audio may be quiet while the main audio channel, at peak deviation, is causing an increase in the SCA noise floor.

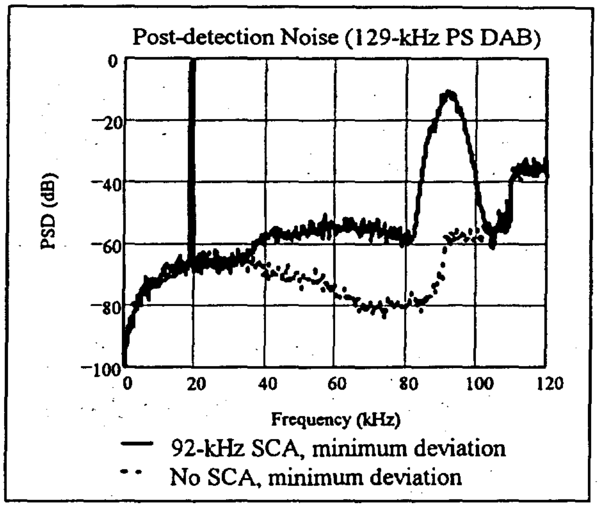

Figure 7. Effects of 92-kHz SCA

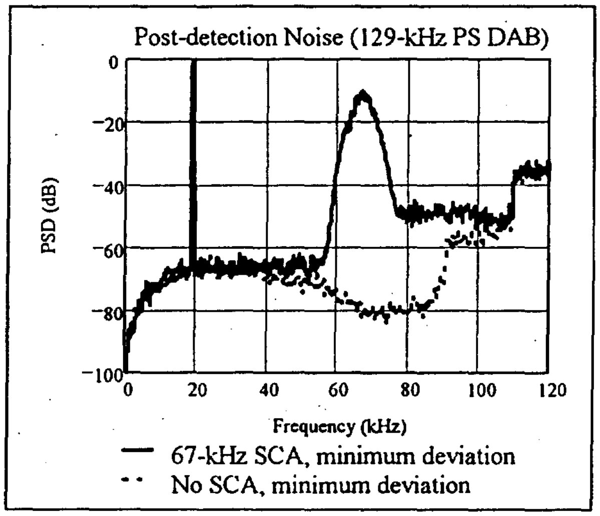

Figure 8. Effects of 67-kHz SCA.

Simulations were performed using SCAs with peak-deviated audio signals in the presence of a -22-dB, 129-kHz pulse-shaped DAB signal. Figure 9 indicates that the SNR of a 67-kHz SCA (in a 10-kHz bandwidth) is 25-30 dB at the transmitter when the main audio channel is near maximum deviation.

For 92-kHz SCAs, the SNR is 20-25 dB, as illustrated in Figure 10.

Without DAB, typical noise floors are roughly 40 dB. The increase in noise floor should not pose a problem for digital SCAs (e.g., Seiko and Radio Broadcast Data System), since they should be robust enough to operate at reasonably low SNRs.

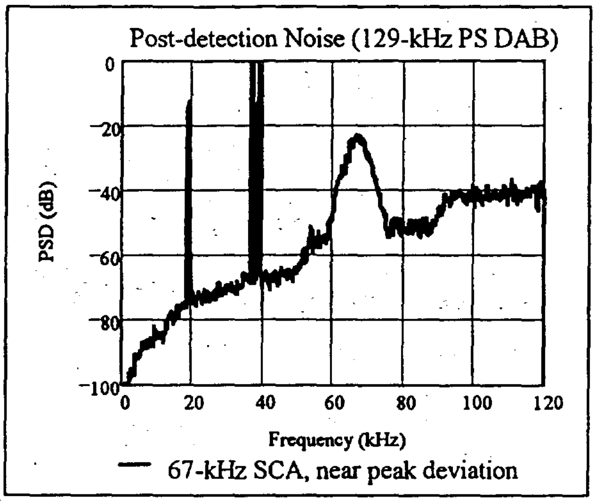

Figure 9. 67-kHz SCA Performance.

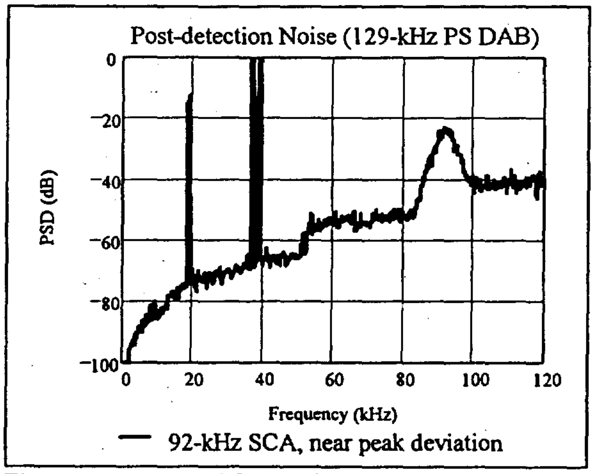

Figure 10. 92-kHz SCA Performance.

Adjacent channel performance

The Federal Communications Commission defines the "edge of coverage" for class B stations as the 54-dBu coverage contour, where 0 dBu is equivalent to 1 microvolt per meter field strength. Class B stations are protected to the 54-dBu contour from 48-dBu interference introduced by first adjacent stations. Thus, at the edge of coverage, the minimum desired-to-undesired signal ratio (D/U) is 6 dB. Simulations have quantified the amount of degradation that would be introduced into the desired IBOC DAB host FM signal when located at the edge of coverage and subject to interference from a -6 dB IBOC DAB first adjacent.

To properly interpret simulation results, it is first necessary to calculate the audio SNR of the simulated receiver at the 54-dBu contour, assuming that the noise contribution is due solely to ambient additive white gaussian noise (AWGN). Assuming a mid-band carrier frequency of 100 MHz and a half-wave dipole antenna, electric field intensity E (V/m) can be converted to carrier power C (W) at the input to the FM receiver using

$$C = \frac{E^2}{120\pi} A_e$$

where $A_e = 1.177\ m^2$ is the effective aperture of the half-wave dipole antenna. Using this formula, a 54 dBu field strength corresponds to a -91.1 dBW carrier power.

An ambient noise temperature of 10,000 K is representative of the FM frequency band; in a 15-kHz bandwidth, this temperature produces a noise power of -146.8 dBW. Hence, a carrier power of -91.1 dBW at the receiver antenna terminals would yield a 55.7 dB/15 kHz carrier-to-noise ratio (CNR). The receiver noise characteristic enables one to determine audio SNR given an input CNR. Using the measured noise characteristic of the simulated FM stereo receiver, this input CNR corresponds to an audio SNR of 64.4 dB/15 kHz.

Recall that a -22-dB pulse-shaped IBOC DAB signal starting at 129 kHz yields an audio SNR of 77.6 dB at the transmitter (and at the edge of coverage if ambient noise were ignored). The preceding calculations demonstrate that, at the edge of coverage, the contribution to audio SNR is dominated by ambient noise; the effects of -22-dB, 129-kHz pulse-shaped IBOC DAB are negligible.

In the simulation, a -22-dB pulse-shaped DAB signal starting at 129 kHz was added to both a quiet FM host at the transmitter (10%-deviated 19-kHz pilot, with no audio or SCAs) and a -6 dB, fully modulated first adjacent. A 150-kHz pre-detection filter (300-kHz total 3-dB bandwidth) was used in the simulated FM stereo receiver. The signal at the input to the FM demodulator is shown in Figure 11.

Results indicate that introduction of the adjacent IBOC DAB channel degrades the audio SNR to 50.0 dB. Although the simulation was performed with the desired signal located at its transmitter, it is clear that the introduction of noise associated with translation to the edge of coverage would be negligible. Therefore, when a -6-dB first-adjacent FM/DAB signal impinges on a quiet host FM/DAB signal at the edge of coverage, the SNR degrades from around 64 dB to 50 dB. Figure 12 illustrates this effect.

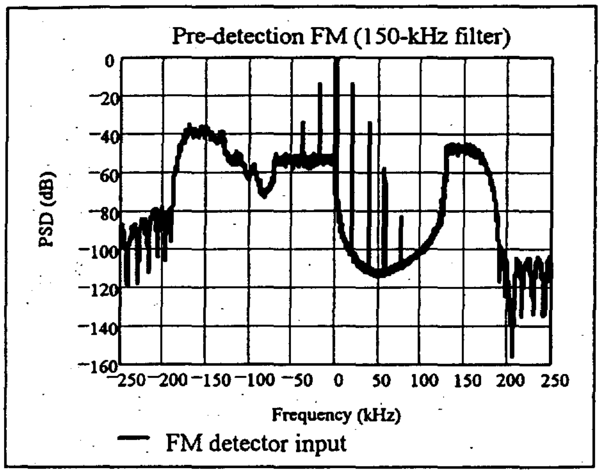

Figure 11. Pre-detection Effect of First Adjacent with 129-kHz PS DAB at edge of coverage.

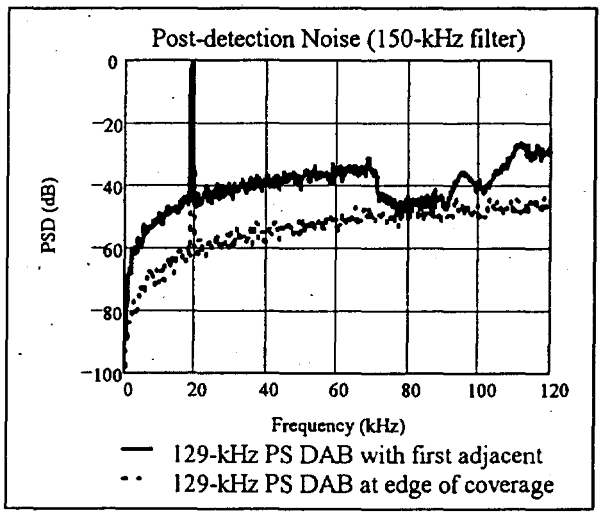

Figure 12. Post-detection Effect of First Adjacent with 129-kHz PS DAB at Edge of Coverage.

While the SNR is diminished, it should be noted that the degradation is highly geographically localized; performance will improve rapidly as the receiver moves farther from the interfering station or closer to the desired station. In addition, due to receiver implementation constraints, actual receivers will experience less than 64-dB SNRs at the edge of coverage. Most automotive receivers are blended to mono at the edge of coverage anyway, mitigating the effects of first adjacent DAB interference. This would improve audio SNR by removing the effects of noise around the stereo subcarrier.

Stereo subcarrier demodulation

During EIA testing of the USADR FM-1 IBOC DAB system, certain inexpensive FM stereo receivers suffered an increase in audio noise when receiving an IBOC DAB FM stereo signal [4]. When the DAB signal was removed from the FM signal, the audio noise disappeared. Investigations revealed that the problem was caused by inadequate filtering of the post-detection baseband stereo multiplex signal. The new baseline DAB waveform has been designed to mitigate this effect.

To recover the stereo information, the 30-kHz-wide, double-sideband amplitude-modulated (DSB) left-minus-right (L-R) signal centered at 38 kHz is demodulated using a 38-kHz local oscillator (LO), and subsequently filtered with a 15-kHz lowpass filter. In most receivers, the 38-kHz LO is simply a square wave, with a 38-kHz fundamental and odd harmonics at 114 kHz, 190 kHz, etc. As a result, in the absence of adequate filtering, not only is the desired L-R signal recovered, but so is any energy in the multiplex signal that lies within ± 15 kHz of 114 kHz and 190 kHz.

In the presence of AWGN only (no DAB), this effect is not pronounced. A well-known property of large-signal FM detection in AWGN indicates that the power spectral density of the post-detection noise is directly proportional to the square of the frequency. Hence, the noise power spectral density at 114 kHz is 9 times that at 38 kHz (9.5 dB), and the noise at 190 kHz has 25 times the power (14.0 dB). High noise levels are mitigated because the amplitude of square wave harmonics decreases with their order: if the 38-kHz fundamental has unit amplitude, the 114-kHz third harmonic has amplitude 1/3 (-9.5 dB), and the 190-kHz fifth harmonic has amplitude 1/5 (-14.0 dB). Therefore, the noise contribution from each harmonic is equal to the noise under the desired signal; this causes a 4.8-dB degradation due to AWGN alone (without DAB) in receivers which do not filter the noise around their LO harmonics.

This decrease in SNR is avoided in well-designed receivers. Some receivers use "Walsh" decoders; others simply filter the baseband multiplex signal prior to DSB demodulation, which effectively eliminates components outside the desired L-R band. Most receivers – even those without such post-detection protection – should ameliorate the effects of the 190-kHz fifth harmonic by pre-detection filtering, since a good design would significantly filter the first adjacent FM signal centered 200-kHz from the desired channel.

Thus, in the presence of AWGN alone, certain inexpensive receivers which employ little or no post-detection protection experience up to a 3-dB stereo SNR degradation (from their DSB LO third harmonic) when compared to their more carefully designed counterparts. Of course, no significant degradation exists when receiving a monaural signal.

As discovered during FM-1 EIA testing, this 3-dB stereo SNR degradation increases when IBOC DAB is added to the analog FM signal [4]. In order to scope the magnitude of the problem, simulations were performed using a well-designed FM stereo receiver with ample protection from 38-kHz harmonics.

Three simulations were run: the first simulated performance in a well-designed receiver by adding AWGN only to a quiet analog FM signal at a level which produced a 64-dB SNR in the left audio receiver channel. The second added DAB only (from 78 kHz to 197 kHz) to the quiet FM signal, at a level which likewise produced a 64-dB SNR in the left audio channel. As shown in Figure 13, the post-detection noise power in the 0-53-kHz audio band is identical for the two simulations (hence the equal 64-dB SNRs).

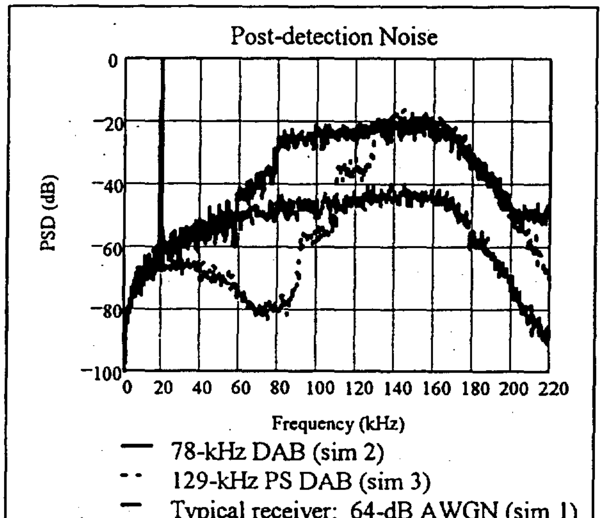

Figure 13. Effect of DAB on 114-kHz Noise Floor.

Note, however, that the noise floors diverge above 60 kHz. In fact, the DAB-induced noise floor is approximately 25 dB higher in the 30-kHz band around 114-kHz. If the simulated receiver did not sufficiently filter the post-detection noise floor, the stereo noise increase would have degraded the audio SNR well below 64 dB in the second simulation.

It has been suggested that simply suppressing DAB energy in the 114±15 kHz band would eliminate the post-detection noise in this region. Due to the non-linear nature of the FM demodulator, this is not entirely the case. Instead, simulations have shown that such a notching of DAB carriers creates a 12-dB improvement across the 30-kHz band around 114 kHz. Thus, simply limiting DAB bandwidth might still cause stereo SNR degradation in radios with inadequate post-detection filtering.

Significant improvements, however, were observed in the third simulation, in which the DAB signal was moved beyond 129 kHz *and* pulse shaping was applied. Pulse shaping causes a significant decrease in the noise floor across much of the post-detection band, as illustrated in Figure 13 (129-kHz PS DAB). Figure 14 provides a magnified view of the 30-kHz region around 114 kHz.

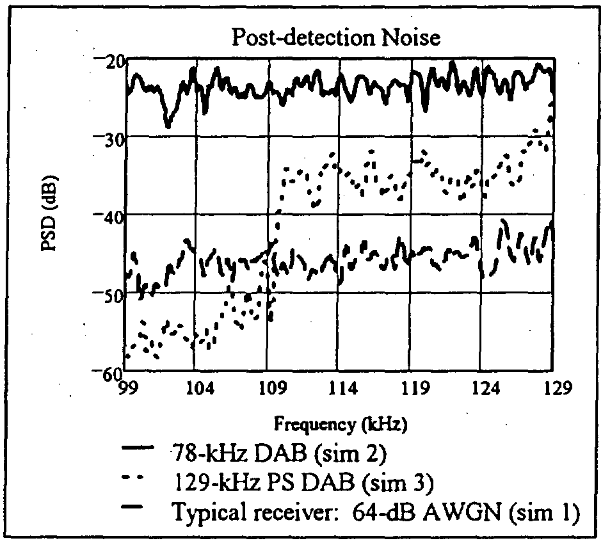

Figure 14. Effect of DAB Placement and Pulse Shaping on 114-kHz Noise Floor.

Note that the noise floor steps up at 110 kHz, due to mixing of the Bessel-weighted 19-kHz pilot harmonics with the DAB signal during FM demodulation. As a result, above 110 kHz, an improvement of around 10 dB is gained over that afforded by 78-kHz (non-pulse-shaped) DAB. Below 110 kHz, however, a 30-dB improvement is observed. Thus, using DAB placement and pulse shaping, the overall stereo noise increase due to the addition of DAB in radios with inadequate filtering can effectively be limited to acceptable levels.

The preceding analysis presents a worst-case bound on the 114-kHz degradation due to DAB; in reality, even the most poorly designed receivers should provide some degree of pre-detection filtering to mitigate the noise level around 114 kHz. Furthermore, in a typical environment, with practical receiver implementations, it is probable that the degradation will be imperceptible to the listener. Note that none of the car radios employed in the EIA tests exhibited the problem [4].

Interference Summary

Westinghouse has analyzed the impact on performance of the host FM signal in the presence of various IBOC DAB configurations. Simulations and analysis indicate that FM performance is least affected when a pulse-shaped DAB signal is placed between 129 kHz and 197 kHz from the FM carrier. Modulation and coding tradeoffs can be exercised to provide the spectral efficiency required to fit the DAB signal within this bandwidth.

This DAB configuration yields an audio SNR of nearly 78 dB during periods of minimum deviation, with noise during louder passages rendered inaudible to the listener via a masking effect. Even when quiet, noise due to DAB will probably be masked by the noise produced in most typical receivers. SCA interference with the host FM should likewise be inaudible, while the SCAs themselves should perform with SNRs of around 20-30 dB/10 kHz (ample margin for a digital SCA).

When a high-level first adjacent interferer with DAB is present, the audio SNR of the host (with DAB) degrades to 50 dB at the edge of coverage. However, the degradation is highly localized. Most automotive receivers are blended to mono at the edge of coverage anyway, mitigating the effects of first adjacent DAB interference.

Finally, a slight degradation may be observed during stereo subcarrier demodulation in existing inexpensive FM stereo receivers. This degradation, which has not been demonstrated in car radios, may prove imperceptible in typical listening environments.

IV. CHANNEL CODING

Forward error correction and interleaving improve the reliability of the transmitted information. In the presence of adjacent channel interference, the outer OFDM subcarriers are most vulnerable to corruption, and the interference on the upper and lower sidebands is independent. The information, coding and interleaving are specially tailored to deal with this nonuniform interference such that the communication of information is robust. Specifically, this nonuniform interference is the focus here where special coding and error handling results in more robust performance.

The IBOC DAB system will transmit all the digital audio information on each DAB sideband (upper or lower) of the FM carrier. Recall that the baseline system constrains the DAB signal to within 130 kHz to 197 kHz above and below the FM center frequency, as shown in Figure 1. Each sideband can be detected and decoded independently with an FEC coding gain achieved by a rate 4/5 convolutional code on each sideband. This redundancy permits operation on one sideband while the other is corrupted. However, usually both sides are combined to provide additional signal power and coding gain. Furthermore special techniques are employed to demodulate and separate strong first adjacent interferers such that a "recovered" DAB sideband can supplement the opposite sideband to improve coding gain and signal power over any one sideband.

The goal here is to transmit the DAB signal on both the upper and lower sidebands such that the sidebands can be independently detected and decoded, each with some FEC coding gain. Additional coding gain, along with some power gain of course, is desired when both sidebands can be combined. The reason for these requirements is that the interference on each sideband is independent of the other; however, the level of interference across the subcarriers on any one sideband is related to the power spectral density of the adjacent interferer. Therefore the grouping of independently detectable and decodable sidebands is appropriate.

In order to effectively achieve coding gain when the pair of sidebands is combined, the code on each sideband should consist of a subset of a larger (lower rate) code. Each subset can be designed through "complementary" puncturing of the lower rate code.

A simple way of constructing a code for this application is to start with a rate 1/3 convolutional code. This code can be generated as shown in Figure 15.

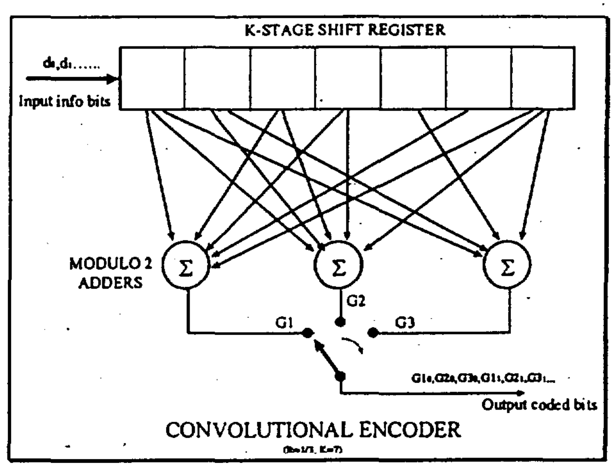

Figure 15. Example of a R=1/3, K=7 convolutional encoder.

The R=1/3 convolutional encoder of Figure 15 can be viewed as producing 3 encoded bit streams (G1, G2 and G3), each at the same bit rate as the input. The combination of these 3 bit streams produces the R=1/3 code output. To create the complementary code pair, for example, a subset of the output code bits is assigned to the lower DAB sideband and a different subset is assigned to the upper sideband. Each subset must contain at least the same rate of bits as the information input rate, plus some additional bits to provide some coding gain. A R=4/5 code on each sideband requires 25% additional bits. One method of allotting bits to the sidebands is presented in Figure 16. The Figure shows the relative spectral locations of the coded bits. These spectral locations are maintained after interleaving by channelizing the interleaver into 8 distinct segments and mapping the segment outputs to the appropriate subcarriers on each sideband.

AN ASSIGNMENT OF CODED BIT STREAMS
TO DAB SIDEBANDS AND SUBCARRIERS

LOWER DAB SIDEBAND		UPPER DAB SIDEBAND	
$G1_0$ $G1_1$ $G1_2$ $G1_3$	$G3_0$	$G3_1$	$G2_3$ $G2_2$ $G2_1$ $G2_0$

G1, G2 and G3 represent generator polynomials (register tap connections) for the R=1/3 code

Figure 16. Example of bit partitioning.

V. BLEND WITH TIME DIVERSITY

Perhaps the most effective method for dealing with the nonstationary mobile radio channel is to provide time diversity between two independent transmissions of the same audio source. Both AM and FM IBOC DAB concepts inherently provide this ability by delaying the analog transmission by a fixed time offset relative to the decoded DAB audio transmission. When the DAB transmission is blocked (or corrupted for any reason) for a short time, then the outage at the DAB decoder is heard after the diversity delay. This diversity delay is incurred at the receiver and is comprised of deinterleaving and FEC decoding delay, audio decoding delay, and any additional delay for diversity improvement. The FEC decoder can be use to identify faulty audio frames and, therefore, the exact time of the DAB audio outage can be predicted. If the channel becomes unblocked after the diversity delay, then the analog signal can be demodulated such that its detected audio output can be blended in while blending out the faulty DAB segment. The listener may detect the temporary degradation in audio quality during the analog blend duration, but will not experience muting or undesirable artifacts.

If the diversity delay is sufficiently large such that the DAB and analog outages are independent, then the probability of an outage after diversity is the square of the probability of outage without diversity. For instance, if the probability of an outage is 1.0%, then the probability of outage after diversity is 0.01%, which is a great improvement. The actual performance can be quantified with knowledge of the autocorrelation function of the channel outage due to severe impairment. This autocorrelation function is expressed as

$$R(\tau) = E\{x(t) \cdot x(t-\tau)\}$$

where $x(t)$ is defined as the stochastic process of the channel loss probability such that a "1" is assigned when the channel is lost and a "0" is assigned when the channel is clear, and τ is the diversity delay time offset. The probability of outage without diversity is $p = E\{x(t)\}$. The autocorrelation function represents the probability of channel outage after diversity improvement as a function of time offset.

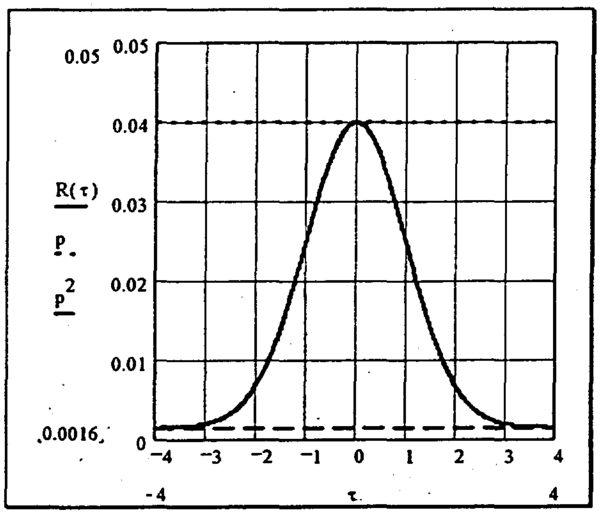

Figure 17. Example Autocorrelation Function of channel loss due to blockage or severe impairment (p=0.04).

An example autocorrelation function is shown in Figure 17; however, an actual autocorrelation function depends on distance from the station, terrain, propagation conditions etc. The figure shows that if the analog signal is not delayed relative to the DAB signal (zero time delay), then the outages are correlated and no benefit is gained from blending since the probability of outage remains the same as without diversity. If the delay is large, then events become uncorrelated and the probability approaches the square of the probability without diversity.

The blend feature also solves the problem of fast tuning time. Without blend, a receiver would incur the diversity delay after tuning to a station before the listener hears the audio. The blend feature will demodulate the analog signal almost instantly, allowing the listener to hear the selection before blending to DAB several seconds later.

VI. ALL-DIGITAL DAB WITH TIME DIVERSITY

The IBOC designs permit evolution to an all-digital DAB format. Without the host analog signal present, the DAB will be transmitted within the primary spectral channel allocation (± 100 kHz). Adjacent channel interference issues are alleviated. The DAB power can be increased by as much as 20 dB, substantially increasing the DAB coverage area. The transmission format will include normal compressed audio plus a more compressed monophonic version of the same signal which is delayed by the diversity time offset. FEC with interleaving is applied to the normal compressed audio. The lower rate audio signal is used for blending during outages in place of the analog signal in the IBOC technique. Furthermore this lower rate audio signal employs FEC coding without interleaving. Therefore the all-DAB signal format facilitates fast tuning and exploits time and frequency diversity with the lower-rate redundant digital audio signal.

VII. AM OFDM IBOC SYSTEM DESCRIPTION

The AM DAB system is only briefly described here. Many of the characteristics of the AM DAB are similar to the FM DAB such as OFDM modulation, audio compression and time diversity. The AM DAB signal is comprised of subcarriers spaced at 500 Hz as illustrated in Figure 18. The symbol rate of each subcarrier is slightly less than 500 Hz due to the guard time between symbol pulses. The carrier complexity starts at BPSK for the 2 carriers closest to the AM carrier, carrying 1 bit per symbol. The subcarriers from 1 kHz to 5 kHz on either side of the AM carrier employ (1,7) modulation carrying 3 bits per symbol while the carriers from 5.5 kHz to 14.5 kHz on either side of the AM carrier employ 32 QAM carrying 5 bits per symbol. Signal processing techniques are employed to reduce the mutual interference between the AM and DAB signals.

The anticipated digital audio rate is 48 kbps, providing high-quality stereo, which is remarkably superior to standard AM. A modest datacasting rate of 2.4 kbps is also provided. The blend-to-analog feature for time diversity is also employed in the AM DAB system to yield robust performance in adverse conditions.

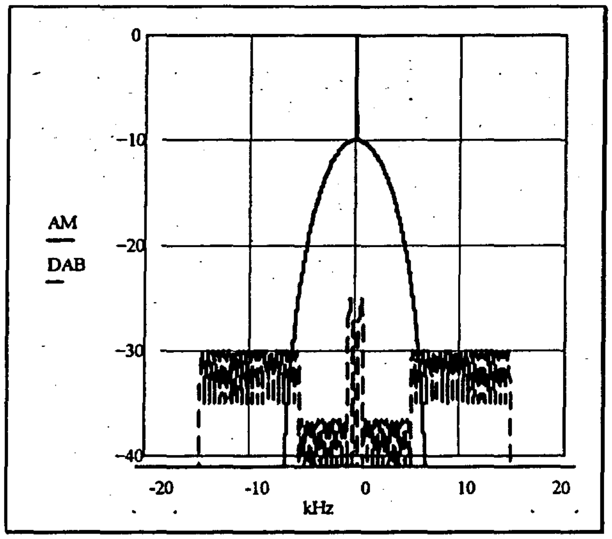

Figure 18. Spectrum of AM plus DAB.

The AM channel is being characterized at 3 frequencies across the AM band. The characterizations consist of transmitting and receiving a flat wideband signal, then analyzing the phase and amplitude characteristics. Channel information is collected only during identified events where the phase and/or amplitude is changing. This data is "played-back" as the channel in a simulator which enables evaluation and refinement of the tracking loops used in channel equalization, AGC, and the phase-locked loop. The statistics of channel events also determine the application and effectiveness of FEC coding and interleaving.

VIII. CONCLUSIONS

AM and FM IBOC DAB systems are being improved and upgraded by *USADR*. Detailed analysis and simulation results support the viability and robustness of these improved systems, of which demonstrations are anticipated in 1997.

Westinghouse has analyzed the impact on performance of the host FM signal in the presence of various IBOC DAB configurations. Simulations and analysis indicate that FM performance is least affected when a pulse-shaped DAB signal is placed between 129 kHz and 197 kHz from the FM carrier. Modulation and coding tradeoffs can be exercised to provide the spectral efficiency required to fit the DAB signal within this bandwidth.

The FM IBOC DAB system will provide virtual-CD quality stereo audio using redundant spectral sidebands to provide frequency diversity and immunity to first adjacent interference. Time diversity is provided through interleaving. A blend-to-analog feature, with time diversity in the order of seconds, permits virtually instant tuning time while filling DAB audio gaps due to blockages or severe impairments. This feature will dramatically improve coverage in areas characterized by intermittent blockages.

AM IBOC DAB will provide stereo audio quality similar to existing FM analog. AM IBOC DAB will exploit interleaving and blending-to-analog with time diversity features similar to FM IBOC DAB.

AM and FM DAB will offer superior DAB coverage through an option to transition, at a future date, to a reduced-quality analog simulcast or to digital only. This option offers an increase in DAB power with the addition of a supplemental DAB transmission consisting of a lower rate compressed audio signal for time diversity reception and nearly instant tuning. This last feature is extremely effective against intermittent blockages and severe impairments, providing performance impossible to otherwise achieve with only frequency diversity, interleaving and FEC.

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[3] Robert D. Culver, "*Report of the Field Test Task Group; Field Test Data Presentation; Working Group B Testing of the CEMA-DAR Subcommittee*; Appendix 6 - FM Modulation Increasing Baseband Noise in the Presence of an IBOC Digital Signal", Lohnes and Culver, Washington, DC, December 1996.

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[5] *Method and System for Simultaneously Broadcasting and Receiving Digital and Analog Signals*, D. Kumar and B. Hunsinger, Patent Application S/N 08/294,140 filed July 1994.

Appendix 13

Summary of
DAR Field Testing Procedures

The complete field testing description, procedures, test routes and data are contained in Reference [2], from which this summary is derived.

San Francisco, California was selected as the test venue for its presentation of varied reception conditions, including urban, rural and mountainous terrain, tall buildings, over-water paths, foliage, rural and mountainous areas, industrial and residential areas, and the availability of suitable transmission site(s).

A mobile measurement vehicle was constructed and equipped to make measurements on six test routes specified in Figures 2A-2F of this Appendix.

Measurements were recorded on computer every 1.171 cm traveled and included RF signal level and subjective assessments of received audio quality/impairment (clear, impaired, or failed), distance traveled, and the passage of pre-scripted landmarks en route. Audio recordings were made (on 8-channel DAT) and included the received DAR system audio, up to two analog FM stations co-located with the DAR transmit antenna (where possible, and using two reference car receivers), and vehicle cockpit audio during the measurements. Video tape recordings (S-VHS) included the route traveled, spectrum analyzers plots showing instantaneous RF signal level and RF spectrum near in frequency to the desired DAR transmission, and time-code. All measurements and recordings were synchronized using SMPTE time-code.

Twenty five test audio program segments were selected and transferred to compact disc with a total play time of just over 60 minutes.

The main transmission facilities were located at Mount Beacon, near Sausalito, California. Supplemental coverage of the Eureka-147/DAB system (using L-band frequencies) was also tested using additional transmission facilities. The VOA/JPL S-band satellite system was uplinked from the JPL laboratory in White Sands, New Mexico to a NASA TDRS Satellite in geo-synchronous orbit over Hawaii (with an approximate 23 degree elevation angle from San Francisco). Specific parameters are shown in Tables 1-3, below.

Data presentations in [2] include instantaneous RF signal level, vehicle velocity, audio events, and distance, landmark-to-landmark over the entire test route, along with a summary of total data records and subsets showing percentage of the route with received audio in clear, impaired or muted (failed) conditions. These are presented for all systems tested on all routes.

Additionally, all routes were measured on the AT&T IBAC FM frequency with the DAR system off-air, to examine the presence of potential co-channel interfering signals.

Further potential processing and presentation of measured data can be made. These might include equating RF values to a reference level, such as power at the input of the device under test, voltage at the antenna element output terminals, as field strength or power density, etc. The RF data points can be averaged over a sliding window to filter out short or longer term variations such as those due to modulation, multipath or short blockage fades. The measured RF signal can be analyzed for indications of multipath propagation and a "rating" applied to route sub segments. The various measured and calculated parameters can be compared with each other; such as comparing events to velocity or events to multipath presence and average longer term RF level. For shared VHF band system, the background RF can be compared to the device under test RF to investigate failures due to insufficient C/I ratio.

TABLE 1

Transmitter parameters for the Eureka 147 DAB SFN installation in San Francisco, July 1996

Parameters\Sites	Mount Beacon		Mount San Bruno		Round Top Mountain
Latitude	37° 51'03" N		37° 41' 15" N		37° 51' 00" N
Longitude	122° 29'51" W		122° 26' 04" W		122° 11' 30" W
Transmitter DAB RMS Power	200 W		200 W		16 W
Horizontal Beamwidth	40°	40°	40°	40°	14°
Azimuth	35°	135°	150°	5°	45°
Transmission Line Loss	1 dB		1.8 dB		1 dB
Power at Splitter Input	159 W		132 W		---
Power Splitter Loss	0.5 dB	0.5 dB	0.5 dB	0.5 dB	---
Splitter Ratio	1/2	1/2	2/3	1/3	---
Power at Antenna Input	71W	71 W	79 W	39W	12.7 W
Antenna Gain	20.4 dBd	20.4 dBd	20.4 dBd	17.4 dBd	18.9 dBd
Maximum ERP	7.8kW	7.8 kW	8.6 kW	2.15kW	1.0 kW

Table 2

AT&T - LUCENT VHF DAR (IBAC) FIELD TEST
BROADCAST STATION KEIA - 96.9 MHz, CHANNEL 245
SAUSALITO, CALIFORNIA

ENGINEERING SPECIFICATIONS AS INSTALLED
(TAKEN FROM FCC FORM 302 LICENSE APPLICATION)

A. Transmitter Site and Control Point

Geographic Coordinates	37° 51' 04" N
(NAD27)	122° 29' 50" W

Mt. Beacon, Wolfback Ridge Road, Sausalito, Marin County, California

B. Equipment

Transmitter	CCA FM35,000DAB(custom)	
Transmission line	Andrew, Types LDF5-50A and HJ5-50A	44 m
Tower	Self-supporting (existing)	40 m
Antenna	Shively, Type 6813-3-SS, 3-bay, nondirectional half wave-spaced	

C. Height

Height of site above mean sea level	338 m
Height of tower above site	40 m
Overall height above mean sea level	378 m
Elevation of average terrain above mean sea level	51 m
Effective height of antenna above site	14 m
Effective height of antenna above mean sea level	352 m
Effective height of antenna above average terrain	301 m

D. Operation

Channel	245
Frequency	96.9 MHz
Transmitter power output (FCC rounding)	6.0 kW
Filter network efficiency	93.3%
Transmission line efficiency	88.5%
Input power to antenna	4.95 kW
Antenna power gain (circularly polarized)	1.01
Effective radiated power (circularly polarized)	5.0 kW

Table 3 LINK BUDGET FOR LINE-OF-SIGHT DIGITAL AUDIO BROADCASTING RECEPTION AT S-BAND (2.05 GHZ)			
AUDIO BIT RATE (Stereo)		160.00	kbps
Satellite transmitter power		7.00	watts
Satellite transmitter power		8.45	dBW
Frequency		2.05	Ghz
Satellite antenna diameter		5.00	m
Satellite antenna gain		38.02	dBi
Satellite antenna beamwidth		2.05	deg
EIRP		46.47	dBW
Satellite Elevation Angle		25.00	deg
Slant Range		39262	km
Free space loss		-190.51	dB
Atmospheric losses		0.25	dB
(Total PFD in 200 kHz BW)		-116.40	dBW/m2
PFD in 4 kHz		-133.39	dBW/m2
Signal at Antenna		-144.29	dBW
Receive Antenna gain		8.00	dBi
Receive Antenna Point Loss		1.00	dB
Received Signal		-137.29	dBW
Antenna Temperature		150	K
Receiver Noise Figure		1.50	dB
Receive System noise temperature		274	K
Receive System G/T (On Antenna Axis)		-16.37	dB/K

C/No		66.94	dBHz
Bit Rate		52.04	dB
Eb/No Available		14.89	dB
Theoretical Eb/No, B.E.R. = 10E-6		3.50	dB
Receiver implementation loss		1.00	dB
Interference degradation		0.50	dB
Receiver Eb/No Requirement		5.00	dB
LINK MARGIN, Beam Center		9.89	dB
LINK MARGIN, Beam Edge		6.89	dB

Electronic Industries Association

NRSC DAB Subcommittee • Field Test Task Group
"Long Path" Test Routes

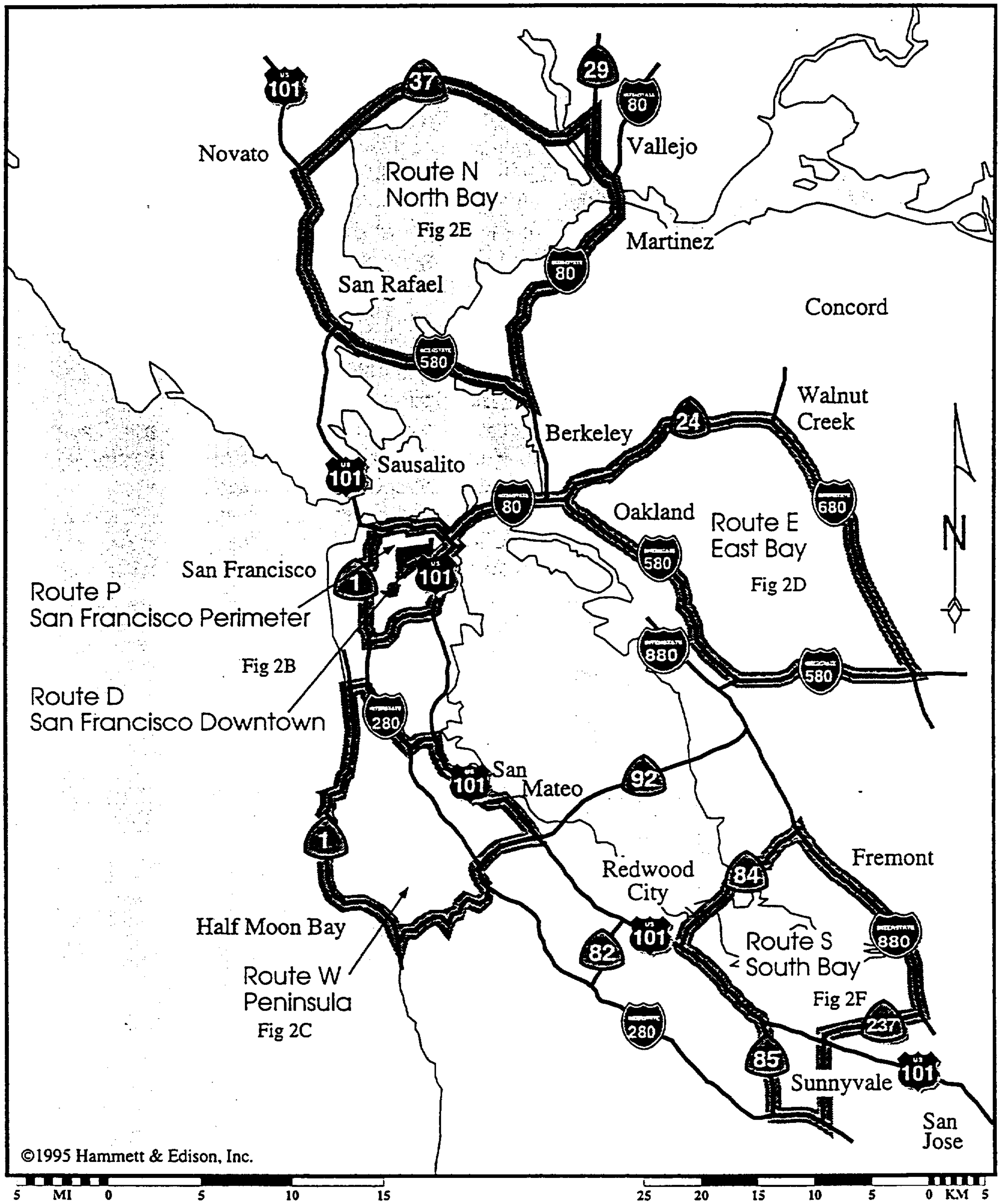

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Figure 2A
Revised June 26, 1996

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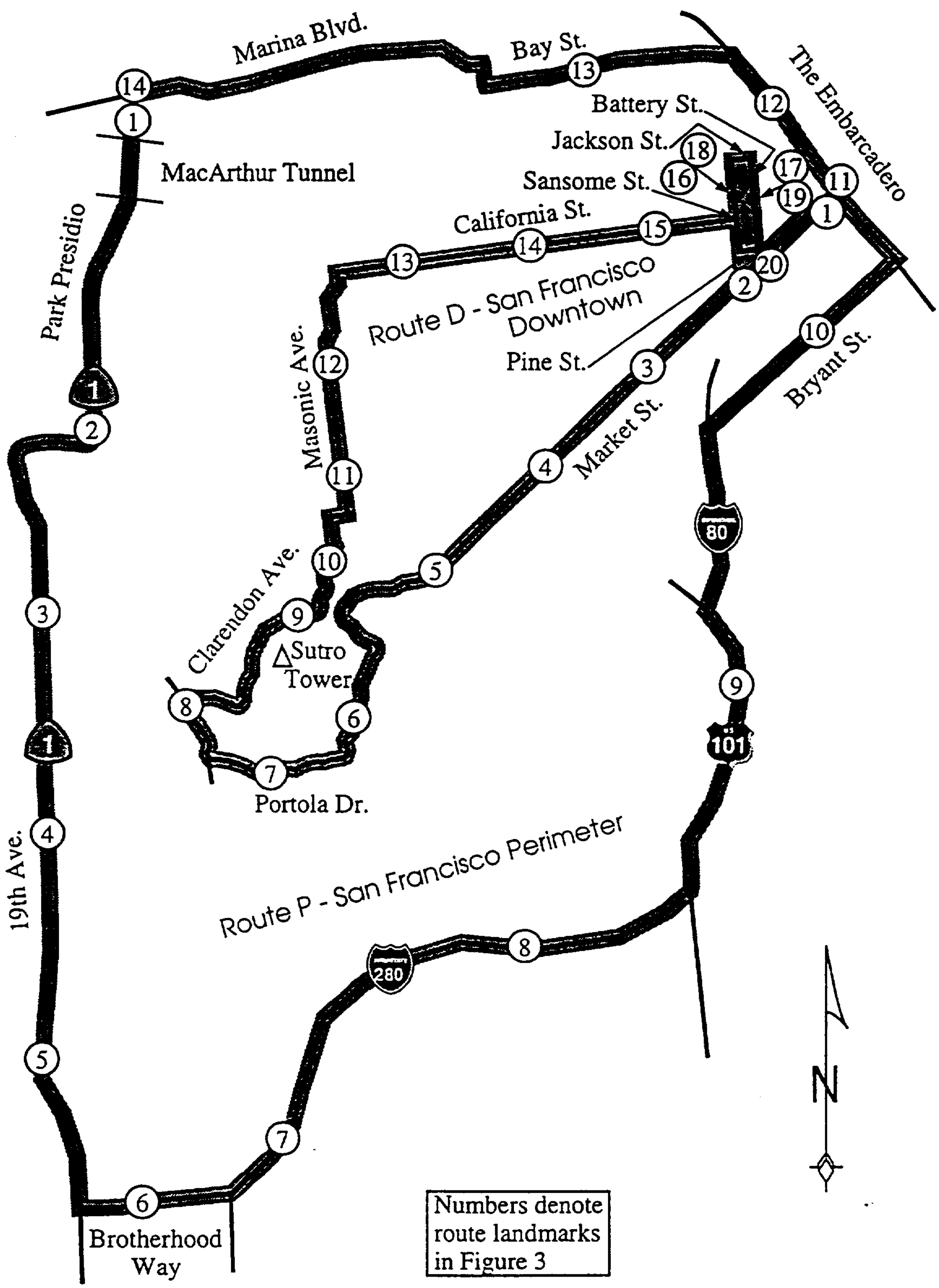
"Long Path" Test Routes

Routes D & P • San Francisco



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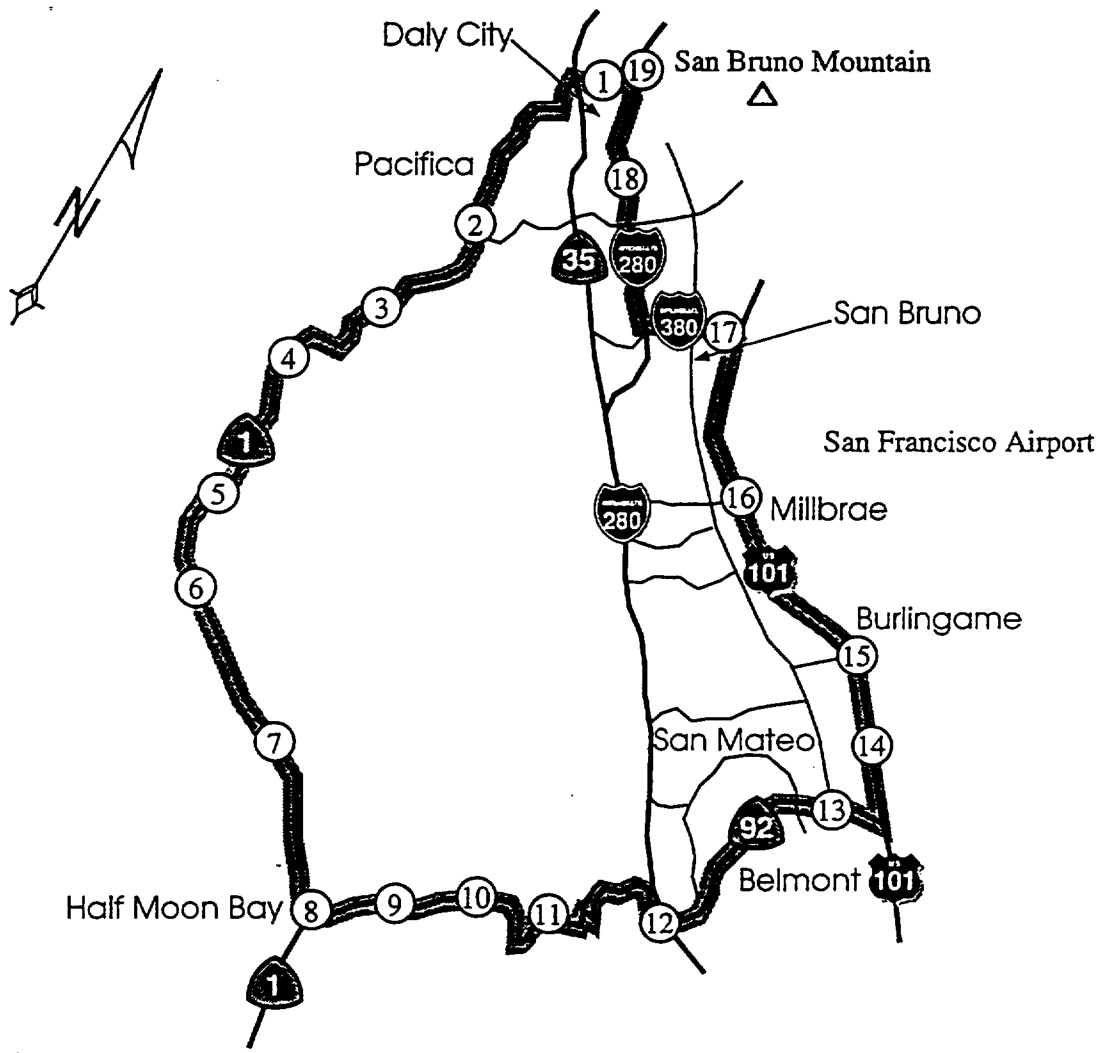
"Long Path" Test Routes

Route W • San Francisco Peninsula



Numbers denote route landmarks in Figure 3

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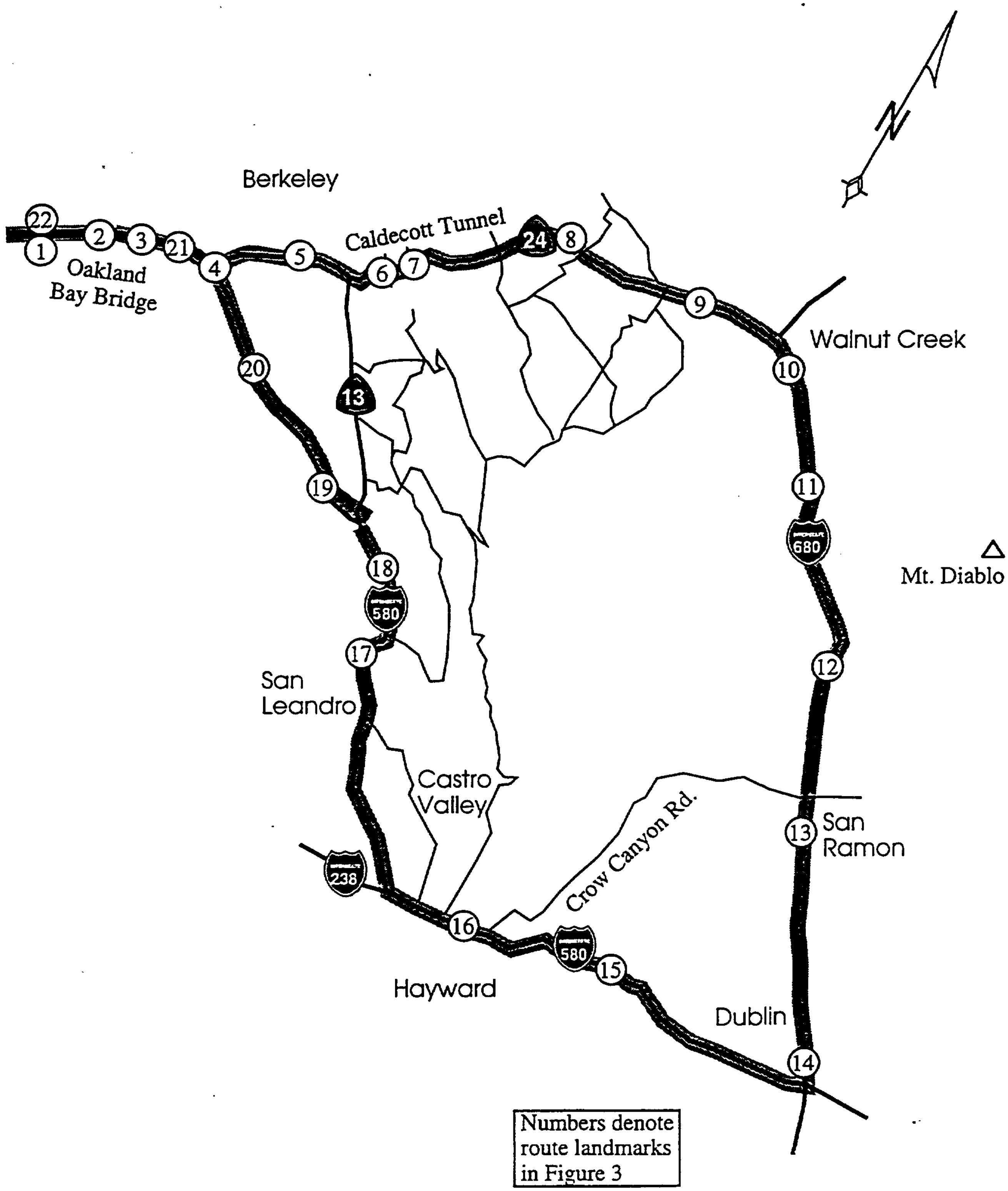
"Long Path" Test Routes

Route E • East Bay



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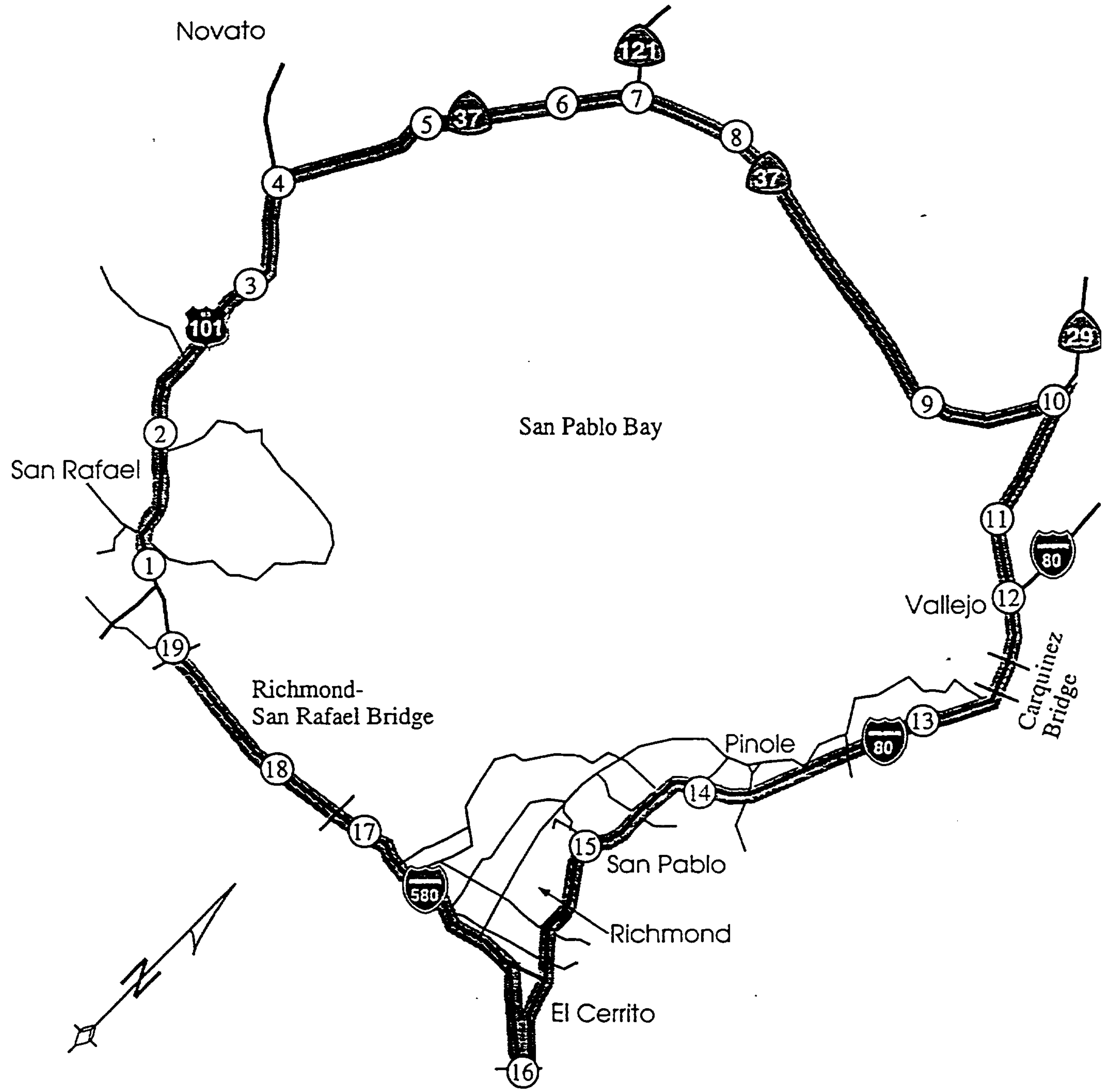
"Long Path" Test Routes

Route N • North Bay



Numbers denote route landmarks in Figure 3

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Figure 2E

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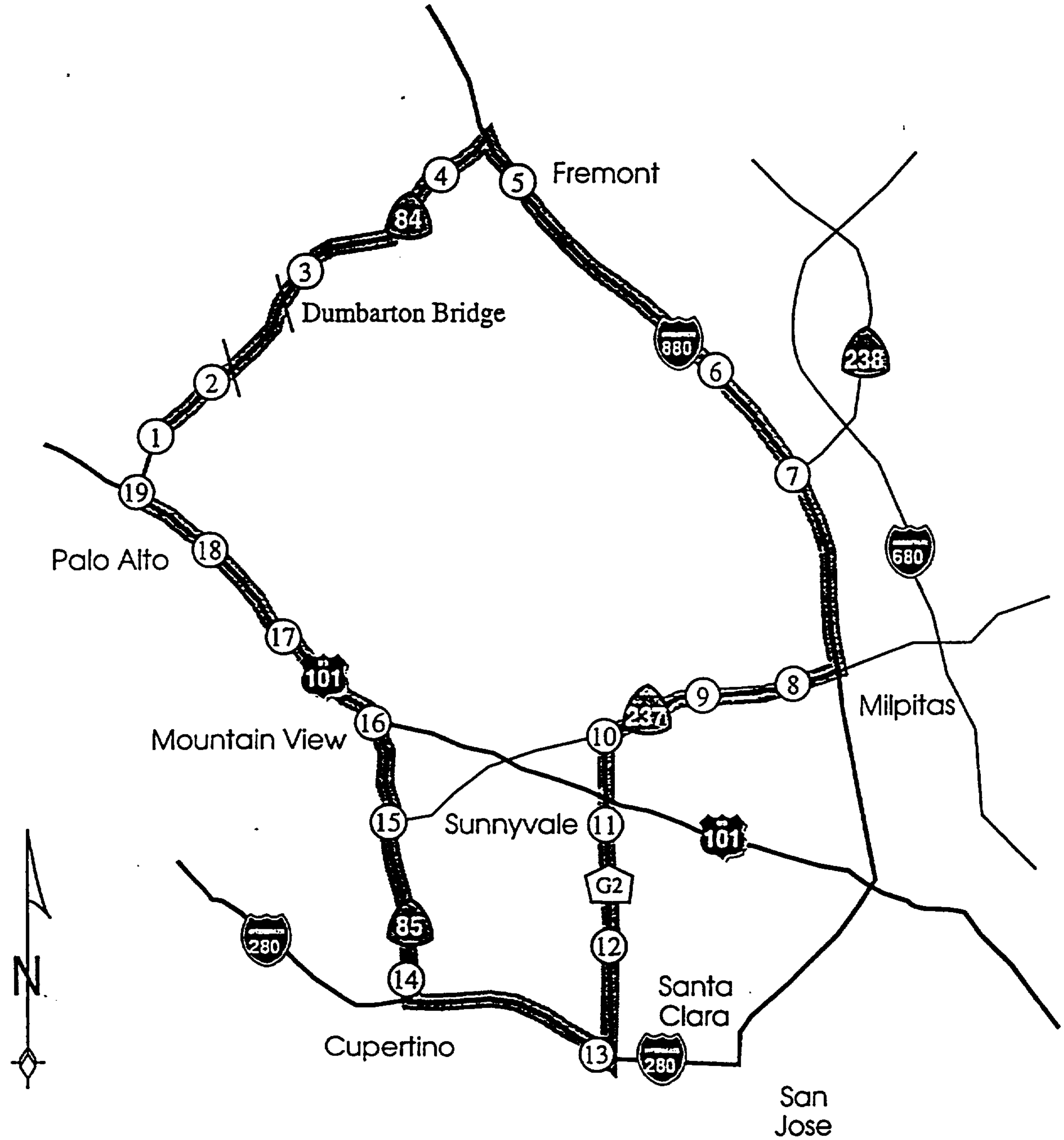
"Long Path" Test Routes

Route S • South Bay



Numbers denote route landmarks in Figure 3

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Figure 2F